Copper FDDI Physical Channel

DOCUMENT IDENTIFIER: A-SP-ELEN529-02-0000 Rev A, 07-Feb-1992

ABSTRACT: This document is a rationale and description of the FDDI (Fiber Distributed Data Interface) copper physical channels. A complete description of the copper distributed data interface physical channels is obtained with the collateral specifications for shielded twisted pair (STP), ThinWire, and unshielded twisted pair (UTP) links.

APPLICABILITY: Copper FDDI applies only to the secondary links in FDDI and not to the main dual counter rotating ring backbone.

STATUS: APPROVED 07-Feb-1992; use VTX SMC for current status.

This document is confidential and proprietary, and is the property of Digital Equipment Corporation. It is an unpublished work protected under applicable copyright laws.

© Digital Equipment Corporation. 1992. All rights reserved.

Copper FDDI Physical Channel

DOCUMENT IDENTIFIER: A-SP-ELEN529-02-0000 Rev A, 07-Feb-1992

REVISION HISTORY:	No previous revision.
Document Management Category:	Product Management Process Requirements (TPR)
Responsible Department:	Network Systems Engineering
Responsible Person:	Simon Ginzburg
Authors:	John Fitzgerald and Simon Ginzburg

APPROVAL: This document has been reviewed and approved.

John Fitzgerald - Network Systems Engineering

Simon Ginzburg - Network Systems Engineering

Direct requests for further information to:

Simon Ginzburg
Use $ VTX ELF for the latest location information.

Use VTX SMC to order copies of this document from Standards and Methods Control. Send distribution questions to JOKUR::SMC or call DTN: 234-4423.

The DIGITAL logo is a trademark of Digital Equipment Corporation.
ThinWire is a trademark of Digital Equipment Corporation.
Teflon is a registered trademark of E.I. du Pont de Nemours & Company.
AT&T is a registered trademark of American Telephone & Telegraph Co.

CONTENTS

Chapter 1 MAJOR CHARACTERISTICS OF COAXIAL AND BALANCED PAIRS 3

1.1 INTRODUCTION 3

1.2 TRANSMISSION LINE MODEL 5

1.3 CHARACTERISTIC IMPEDANCE AND PROPAGATION CONSTANT 6

1.4 PHASE AND GROUP DELAYS—DEFINITIONS 7

1.5 EFFECT OF TEMPERATURE ON THE PRIMARY PARAMETERS 8

1.6 RADIATION 8

1.7 CROSS TALK IN BALANCED PAIRS 9

1.8 REFLECTIONS 10

1.9 IMPULSE NOISE 11

1.10 FREQUENCY RESPONSE 14

1.11 EQUALIZATION 15

1.12 THERMAL NOISE CONTRIBUTION 17

1.13 TRANSMISSION LINK MODEL 18

1.14 TRANSMITTER RISE/FALL TIME 19

1.15 DCD, BASELINE FLUCTUATIONS, AND HYSTERESIS IN THE SIGNAL DETECT CIRCUITRY 19

1.16 COAXIAL CABLES 19

1.17 SHIELDED BALANCED PAIRS 20

1.18 UNSHIELDED TWISTED PAIR (UTP) 28

Chapter 2 COAXIAL LINK 33

2.1 BIDIRECTIONAL TRANSMISSION 33

2.2 REQUIREMENTS TO RETURN LOSS OF JOINTS IN COAXIAL LINE 34

2.3 ACHIEVABLE DISTANCE 35

2.4 CHANNEL STRUCTURE 35

2.5 DATA DEPENDENT JITTER (DDJ) 37

2.6 LINK LOSS DETECTION 37

Chapter 3 SHIELDED TWISTED PAIR (STP) LINK 39
3.1 INTRODUCTION 39
3.2 LINK BLOCK DIAGRAM 39
3.3 LINK DETECTION CIRCUITRY 42
3.4 ACHIEVABLE DISTANCE 43

Chapter 4 UNSHIELDED TWISTED PAIR (UTP) LINK 47
4.1 MULTILEVEL MODULATION 47
4.2 COMPARISON OF DIFFERENT APPROACHES FOR TRANSMITTING FDDI OVER UTP 48
4.2.1 CAP-32 54
4.2.2 PR-4 55
4.2.3 2B1Q 60
4.2.4 MLT-3 and MLT-4 60
4.2.5 3B10 63
4.2.6 QAM-32 63
4.2.7 UTP TRANSCEIVER BLOCK DIAGRAM 64
4.3 ENHANCING TECHNIQUE 64
4.3.1 ADAPTIVE EQUALIZATION 64
4.3.1.1 Zero-Forcing Algorithm (ZFE) 66
4.3.2 Cross Talk Suppressor 71
4.3.3 VITERBI ALGORITHM 72
4.3.4 SCRAMBLING 75
4.3.5 SUMMARY 79

Appendix A REFLECTION LIMITED COAXIAL LINK: ACHIEVABLE DISTANCE 81

Appendix B UTP: ACHIEVABLE DISTANCE 85

Appendix C BIBLIOGRAPHY 87

Appendix D RELATED DOCUMENTS 89
D.1 Other Digital Documents 89

FIGURES

1–1 Star Quad 3
1–2 Twin Quad 4
1–3 Coaxial Cable Diagram 5
1–4 Primary Parameters 5
1–5 Representation of a Lossless cable 7
1–6 FCC Emission Requirements 8

1–7	FCC Emission Requirements	9
1–8	Voltage Standing Wave Ratio (VSWR)	10
1–9	Experimental Data, Impulse Noise	13
1–10	Example of Equalizer Circuit	16
1–11	Transmission Line Model	18
1–12	Cable Types	21
1–13	Type 1 Outdoor	22
1–14	Type 1 Plenum, Non-Plenum, Riser	22
1–15	Type 2 Plenum, Non-Plenum	23
1–16	Type 6 Office	23
1–17	Attenuation and Group Delay, 112 Meters—Type 6	24
1–18	Near End Cross Talk—Type 6	25
1–19	Attenuation and Group Delay, 100 Meters—Type 1	26
1–20	Near End Cross Talk—Type 1	27
1–21	Attenuation Measurements	30
1–22	Near End Cross Talk Measurements	31
2–1	Bidirectional Transmission	34
2–2	Point-to-Point Coaxial Transceiver Link Block Diagram	36
2–3	Coaxial Line	37
2–4	Link Detect DC Circuit Model	38
3–1	Shielded Twisted Pair FDDI	39
3–2	STP Link Functional Diagram	40
3–3	Asymptotic Zero/Pole Approximation to Cable	41
3–4	Cable Detect Implementation Example	42
3–5	Window Opening as a Function of BER	44
3–6	Results of Emission Testing, with FCC Limit	45
4–1	Probability of an Error for Pulse Amplitude Modulation (PAM)	48
4–2	PSD of Copper Distributed Data Interface Line Codes Without Filtering	49
4–3	FDDI UTP Spectrum	50
4–4	Original FDDI Spectral Width	51
4–5	FDDI UTP Spectrum	52
4–6	Prospective Line Codes in the Time Domain	53
4–7	CAP-32 Transceiver Structure	54
4–8	MDB Encoder Circuit	55
4–9	Modified Duobinary (MDB) Format	56
4–10	State Diagram for MDB	57
4–11	Memory Effect	58
4–12	FDDI Line States—Halt (H)	59
4–13	FDDI Line States—(MLS)	59
4–14	FDDI Line States—Idle	59
4–15	FDDI Line States—(J,K)	59
4–16	MLT-3 and MLT-4 Formats	61
4–17	MLT Spectrum	62
4–18	Error Multiplication in MLT-3	63
4–19	Possible UTP Transceiver Block Diagram	65
4–20	Adaptive Filter-Equalizer	66

4–21 Zero-Forcing Algorithm 67
4–22 MSE Equalizer 68
4–23 Decision Feedback Equalizer 70
4–24 Cross Talk Suppressor 71
4–25 Cross Talk Wave Forms 72
4–26 Viterbi Algorithm 73
4–27 Tree Diagram 74
4–28 General Basic Scrambler 75
4–29 Single-Counter Descrambler 77
4–30 Single-Counter Scrambler 78

INTRODUCTION

High costs associated with implementing FDDI are the major drawbacks to its rapid growth. Fiber-optic FDDI networks and adapters are presently priced beyond what some network professionals are prepared to pay. Today, FDDI ports are about 10 times more expensive than Ethernet ports. But the prices of these are expected to drop with second and third generation hardware implementations.

Another reason for FDDI's initial cost is rewiring. Most buildings are not wired for fiber, which can be expensive to install—roughly $770 to go from the wiring closet to the desktop [1][1] . This amount includes $110 for the fiber, $340 for installation, and $320 for connectors and jumpers. Understandably, many network managers are reluctant to rewire their buildings.

Building FDDI networks with coaxial or balanced pairs is a more cost-effective approach. By making FDDI available over existing twisted-pair cabling, the need to rewire fiber at the station level will be eliminated, as will much of the cost associated with fiber-based node adapters. FDDI will encounter less resistance in the marketplace, thereby boosting sales considerably at the high-volume station level and fueling the overall FDDI market.

For the short term, a compromise solution that combines fiber and copper in the same FDDI network is beneficial,—fiber is advantageously used in the backbone and copper at the station level, where it is often already installed. The result is FDDI data rates are much lower in cost.

In other words, copper is chiefly in hub-based networks, where it supplies the lobe (hub-to-workstation) connection. Fiber is the main choice for inter-hub paths. Copper FDDI applies only to the secondary links in FDDI, and not the main dual-counter rotating ring backbone. Secondary links that can be used to create cascaded tree structures that attach to the two main rings provide a single attachment (class B) interface for workstations, lower-level concentrators, and the like. A copper-capable FDDI concentrator will have the conventional dual-attached fiber link if attaching to the main ring, and a set of single attachment ports equipped with copper transceivers. The copper interface in place of fiber optics is the major difference from traditional FDDI network.

This document is a rationale and description of the FDDI copper physical channels. It contains descriptions of ThinWire and shielded twisted pair versions, and a review and comparison of existing approaches to unshielded twisted pair FDDI physical channels.

[1] Throughout this document, numbers that appear within [ ] refer to the numbered references in the Bibliography.

CHAPTER 1

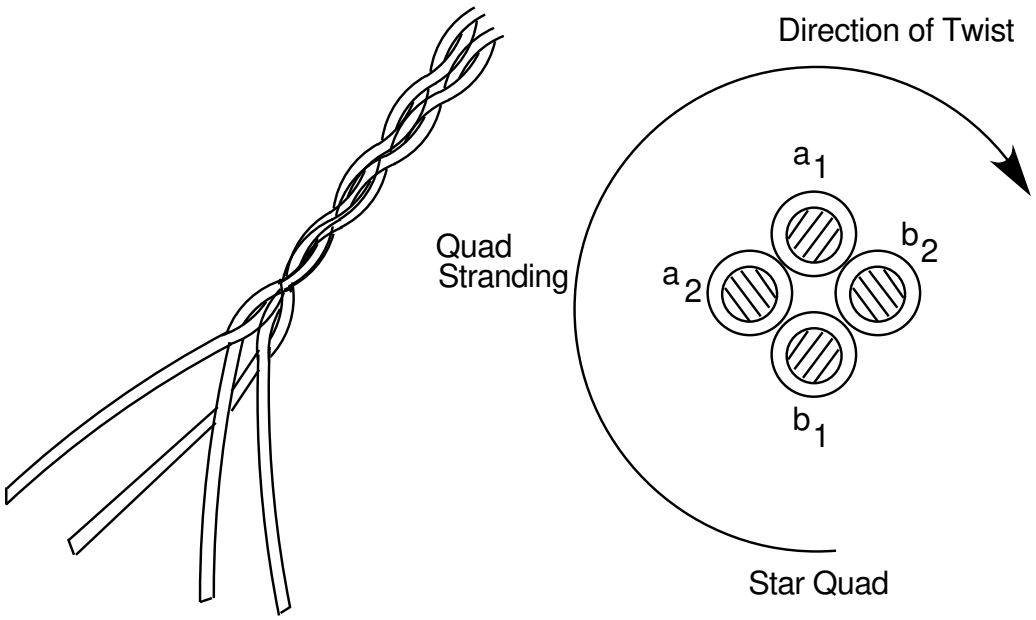
MAJOR CHARACTERISTICS OF COAXIAL AND BALANCED PAIRS

1.1 INTRODUCTION

A balanced pair is a two-wire line in which the two conductors are identical, and have the same capacitance and the same leakage with respect to ground.

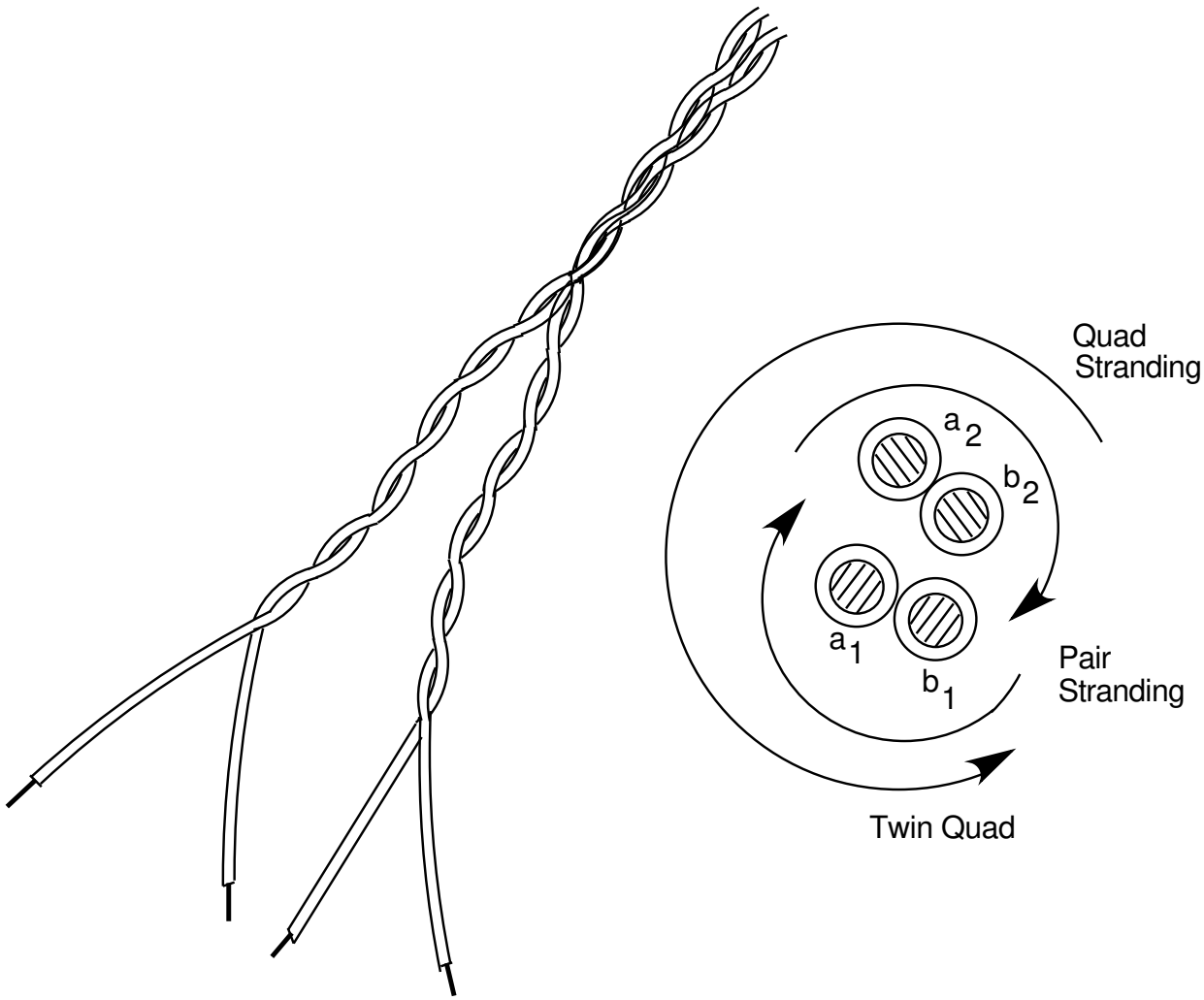
A balanced pair consists of two insulated and twisted metallic wires that are stranded (twisted) in such a way as to reduce the inductive coupling between pairs and reduce the cross talk. The four wires, corresponding to two pairs, can be stranded together to form a star quad. Refer to Figure 1-1.

Figure 1–1: Star Quad



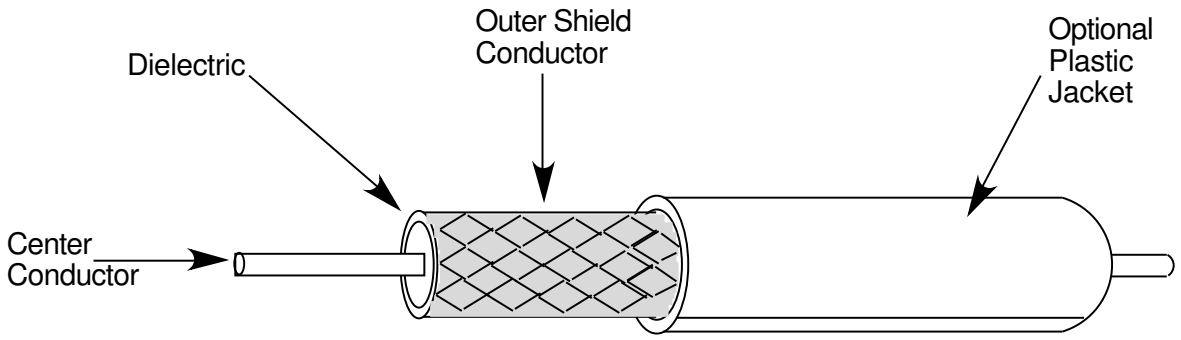
A twin quad is created by stranding two twisted pairs with different twist rates which provides better cross talk properties but requires more space. See Figure 1-2. The whole arrangement is protected from external influences by an insulating jacket. Pairs may have individual or overall shields.

Figure 1–2: Twin Quad



A coaxial pair is imbalanced and consists of two concentric conductors. Refer to Figure 1-3. The exterior conductor is grounded, while the interior conductor is insulated and centered by a dielectric material. The total cable consists of one or several pairs and is protected from exterior influences.

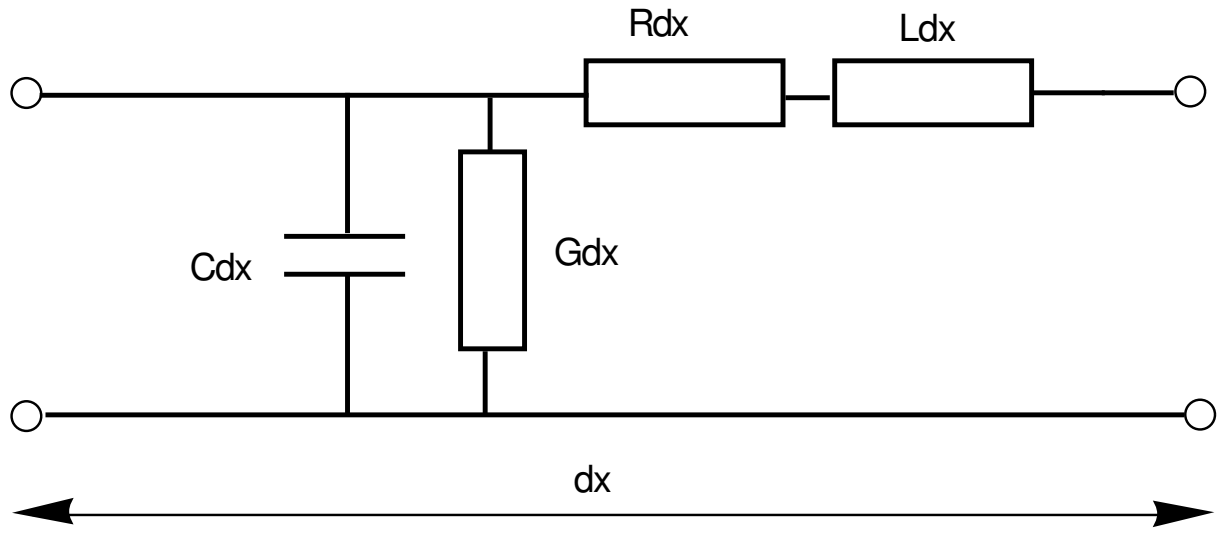
Figure 1–3: Coaxial Cable Diagram



1.2 TRANSMISSION LINE MODEL

Common practice [2, 3] is to represent a transmission line as a very large number of lumped element filters as shown in Figure 1-4. The figure shows the relationship of the geometry of the cable to the distributed inductance (L), the capacitance (C), the resistance (R) and the conductance (G), which are the *primary parameters* of any transmission line.

Figure 1–4: Primary Parameters



- R (*Ohm/unit length*) is the resistance per unit length, which expresses the series resistance of the two conductors per unit length (in practice, generally for 1 km of the line.

For resistance R, at high frequencies, the current in a conductor is largely confined to a small distance of the surface of a conductor (the skin depth). Moreover, this depth changes with frequency—the result being that the skin depth decreases while R increases with frequency.

- L ($Henry/unit\ length$) is the inductance per unit length. One part of it is caused by the magnetic field inside the conductors, dependent on the skin effect, and the other is caused by the magnetic field between the conductors.
- C ($Farad/unit\ length$) is the capacitance per unit length, related to the permittivity of the dielectric located between the conductors.
- G ($Mho/unit\ length$) is the leakage conductance per unit length whose origin is mainly due to dielectric losses (or insulation defects) that increase with frequency.

With reasonable dielectrics at higher frequencies, the effects of G are usually smaller than those of the resistance R.

1.3 CHARACTERISTIC IMPEDANCE AND PROPAGATION CONSTANT

In terms of the primary parameters (lumped elements), the characteristic impedance and propagation coefficient are defined as:

$$Z_0 = [(R + jwL)/(G + jwC)]^{1/2}$$

and

$$\gamma = [(R + jwL)(G + jwC)]^{1/2} = \alpha + j\beta$$

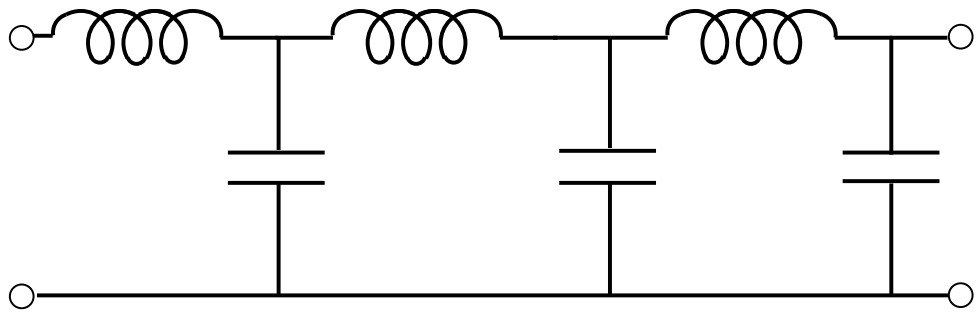
The characteristic impedance Z_0 would be measured at the input of infinitely long line or of a finite line terminated with the same Z_0.

The propagation coefficient consists of a real part—the attenuation coefficient α—and the imaginary part—the phase coefficient β.

Because the primary parameters are functions of frequency, the characteristic impedance and the propagation constant are also functions of frequency.

In most applications when loads are added, the cable can be considered lossless as shown in Figure 1-5. If an infinitely long cable is assumed, then the signal propagating down this cable would suffer no reflections because each segment looks identical.

Figure 1–5: Representation of a Lossless cable



If the ratio of the signal voltage to the current flowing in the conductor is measured at any point it would be found to be a constant. This constant is termed the characteristic impedance Z_0 of the cable. For the lossless case, this impedance is related to the lumped parameter model by:

$$Z_0 = (L/C)^{1/2}$$

1.4 PHASE AND GROUP DELAYS—DEFINITIONS

During a time interval dt, the phase Q of a sinusoidal signal at a point x of a line varies by the amount of

$$dQ = 2\pi\ dt/T = 2\pi * dtf = dt * w$$

where T = 1/f is a period.

The same phase change is established simultaneously between point x and point x + dx such that

$$dx = dQ/b = (dt) * w/b$$

For the speed of propagation or phase velocity V is the ratio:

$$V = dx/dt = w/b\ *\ (m/s)$$

Here b is the phase change coefficient.

Its inverse is called the phase delay per unit length

$$Tq = 1/V = b/w\ (s/m)$$

To provide a situation with no phase distortions, phase delay (measured in cycles) versus frequency should be a linear function of frequency. In reality it is not.

An envelope delay is defined as the slope of such curves and may be measured in seconds. In effect, envelope delay at a given frequency is the delay that would be suffered by a very narrow bandwidth signal transmitted at that frequency.

A similar characteristic is called a group delay per unit length where

$$Tl = dB/dw\ (s/m)$$

The propagation times are obtained by multiplying these delays by the length of the line.

1.5 EFFECT OF TEMPERATURE ON THE PRIMARY PARAMETERS

The primary effect of temperature on the transmission characteristics is caused by the variation in the resistivity of the wire. With a continuous current, R varies proportionally with the resistivity. For example, it increases by approximately 4% per degree (resistivity is measured by n ohm/m at dc current).

At high frequencies, due to the skin effect, R increases only proportionally to the square root of the resistivity, approximately 2% per degree [2].

1.6 RADIATION

The Federal Communication Commission (FCC) imposes a set of emission requirements for data communication networks. For UTP, which for the FDDI spectrum, turns into a broadcasting antenna, these requirements on radiation limit the allowable driving voltage. Figures 1-6 and 1-7 show the FCC emission requirements as a function of frequency.

Figure 1–6: FCC Emission Requirements

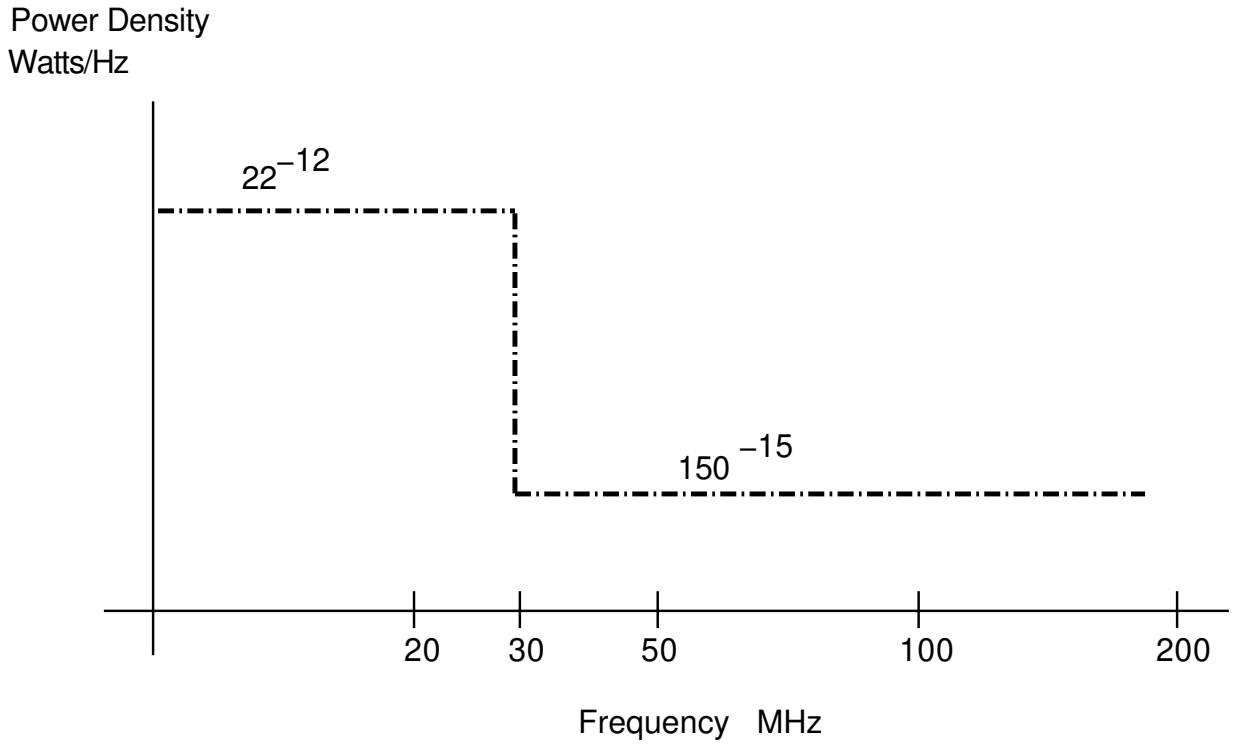
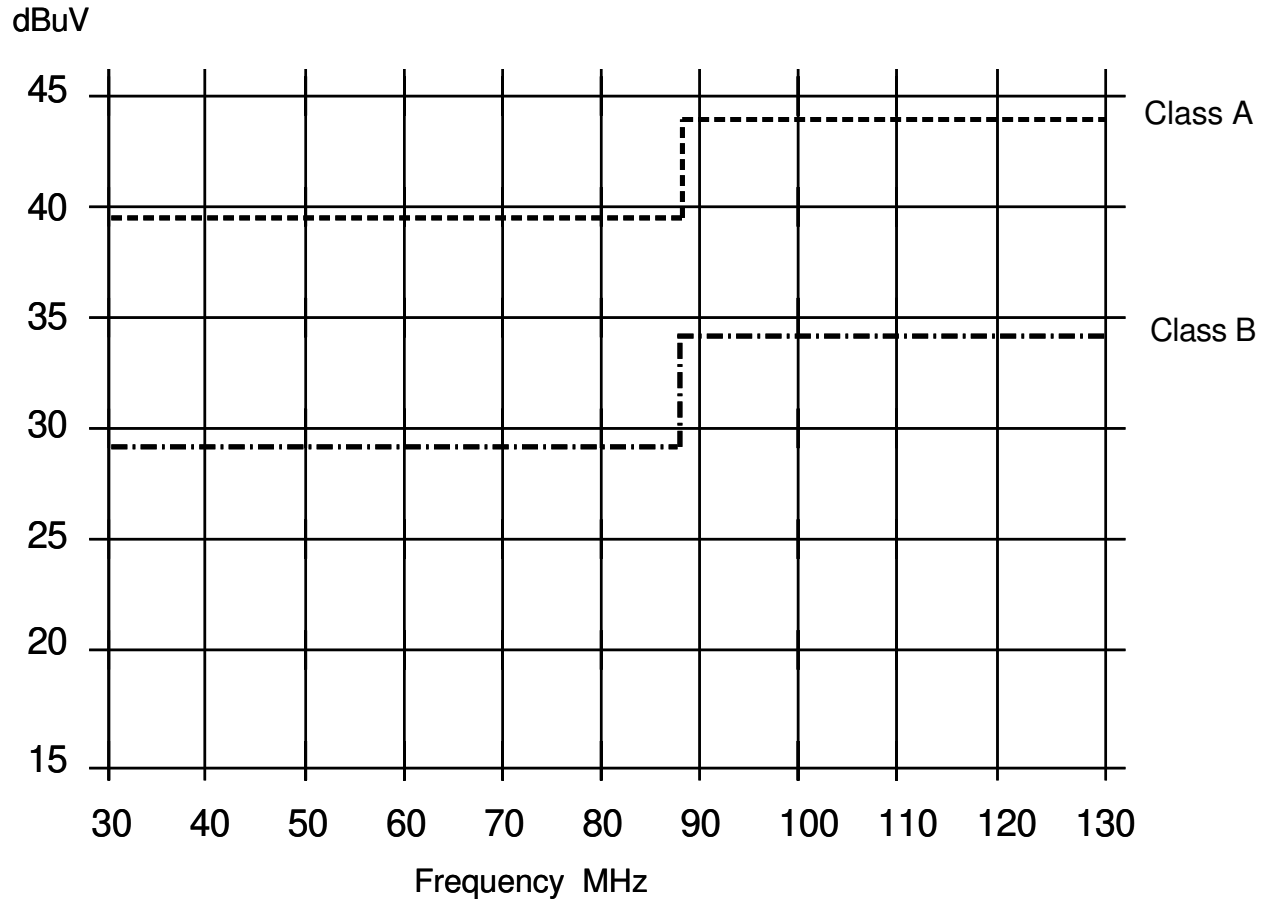


Figure 1–7: FCC Emission Requirements



1.7 CROSS TALK IN BALANCED PAIRS

When two lines are very close (for example, two pairs situated within the same cable sheath), they exert a mutual influence on each other caused by three types of coupling:

1. Galvanic coupling due to a common resistance of the two different lines. This phenomenon notably appears in the case of single-wire lines with a common ground.
2. Capacitive coupling due to the presence of capacitances between the conductors of the two lines.
3. Inductive coupling due to the magnetic field on one of the lines crossing the other (mutual inductance).

These couplings depend essentially on the geometric configuration of the conductors, particulary their proximity to each other. As a result of these couplings, part of the signal transmitted by one line appears on the other and vice versa. This cross talk is highly undesirable since it usually interferes with the signal to be received.

The cross talk level between balanced pairs of the same cable depends greatly on the relative position of the two pairs and the frequencies considered. In the case of near end cross talk, phase variations of different couplings along the line as a function of distance and frequency, provoke great fluctuations in the cross talk attenuation.

When there are many independent disturbers, such as in a 24-pair UTP cable, the cross talk is described by truncated Gaussian distribution [4]. The truncation value is considered as peak cross talk value.

The cross talk usually directly effects the achievable distance by reducing the signal budget.

1.8 REFLECTIONS

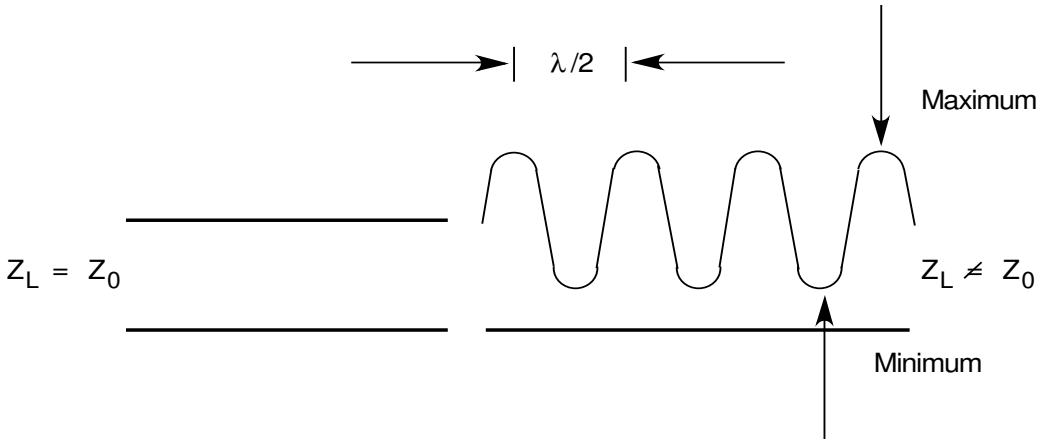
Reflections are caused by improper terminations. The reflection coefficient is defined as

$$m = \frac{(Z_1 - Z_0)}{(Z_1 + Z_0)}$$

The reflection coefficient can be positive, negative or complex, depending on values of characteristic impedance Zo and the termination impedance Z.

Whenever the cable is interrupted or has a disruption in uniformity (bent, twisted, crushed, and so forth), there is the possibility of an impedance mismatch at the new transition. This mismatch will cause a reflected signal. This signal added to the primary incoming signal creates a standing wave pattern that can be characterized through a parameter called the voltage standing wave ratio (VSWR), as shown in Figure 1-8.

Figure 1–8: Voltage Standing Wave Ratio (VSWR)



In equation form:

$$VSWR = (Emax/Emin) = (Ei + Er)/(Ei - Er) = (1 + |m|)/(1 - |m|)$$

where
Ei and Er are the incident and reflected signal amplitude and m is the complex-valued reflection coefficient.

The spatial separation of maximum and minimum is determined by the frequency of voltage source and the associated wavelength on the transmission line, while the position of maxima and minima determined by the phase of the load impedance and the frequency of the source voltage.

The magnitude of the reflection coefficient varies from 0 to 1 and the VSWR from 1 to infinity. If the impedance of the added device is equal to the characteristic cable impedance, maximum power will be transferred to the device and no reflection will occur (VSWR = 1). This condition is known as matched impedance.

Clearly, this transfer is the preferable situation, and designers try to produce devices that result in a VSWR as close to 1 as possible.

An alternative parameter that is often used in specifying the mismatch conditions is the return loss. This mismatch is defined as the ratio of the incident signal amplitude to the reflected amplitude expressed in decibels (dB). The equation is given by:

$$Return\ loss = 20log(Ei/Er) = 20log(1/|m|)$$

Clearly a specification of VSWR, reflection coefficient m, or return loss are equivalent in terms of information content. Each of these parameters is a way to represent impedance mismatching at any point on a transmission line.

A single pulse applied to a line (TDR method) will give rise to multiple reflections traveling back and forth if the line does not have proper matching terminations on both ends.

As mentioned, a reflection coefficient from a cable junction will be a function of frequency, because Zo depends on frequency. The actual reflected pulse train will be a result of averaging over all the frequencies in the signal spectrum.

If a network analyzer is used to evaluate transmission characteristics of the FDDI signals, the result should be averaged over all FDDI frequencies. Contrary to a network analyzer, a TDR shows the average results.

1.9 IMPULSE NOISE

Impulse noise is noncontinuous noise that consists of irregular pulses or noise spikes of short duration and relatively high amplitude.

The major causes of impulse noise are sharp voltage changes in adjacent wires or equipment induced noise spikes in the communication channel.

Examples are:

- Any switches or relays that make or break circuits are the potential sources of impulse noise.
- Power supply induces hum or high frequency components.
- Lightning flashes, radar transmitters, trains, electrical machinery.
- Circuit faults, such as poor quality soldering, dirty relay contacts, or jacks.
- Nonsoldered twisted joints may cause noise due to changes in temperature and slight movements of the joint.
- Elevator shafts

- X-ray equipment

The most likely result of a long error burst is that a large number of data symbol violations will occur.

Due to impulse noise, the FDDI ring may

a. Be disrupted, or

b. The received frame is apparently correct, but is actually errored [6].

Disruption of the FDDI ring includes lost token. The ring requires re-initializiation (entering the claim token process) [10]. Another type of disruption is when an error occurs in a frame and the frame is discarded because the FCS is no longer correct or perhaps the frame delimiter has been altered. The most likely result of this type of disruption is that a higher layer software entity in the transport layer will re-transmit the frame.

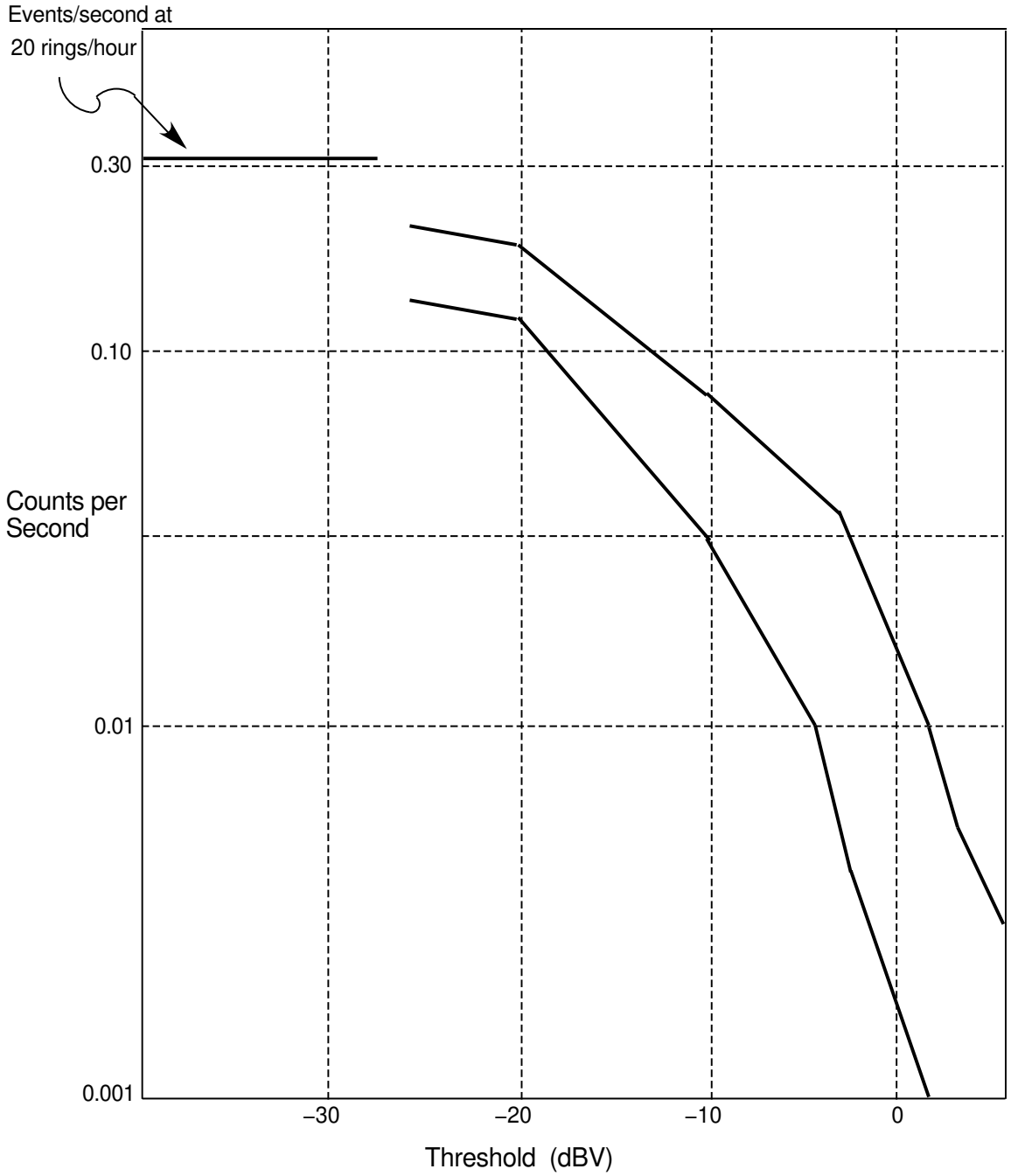
Another possibility is that an error occurred, but the result is a frame that is apparently correct, but in actuality, is not the same as the frame that was sent.

The duration of the impulse noise can be very short or it may last in the order of milliseconds. Therefore, any BER (bit error rate) type characteristics depend on amplitude-duration (two-way) distribution for the impulse noise. This characteristic must be determined for the particular channel bandwidth employed.

Some estimations can be done, based on one-way distribution.

Experimental data of the impulse noise, caused by ringer cross talk for UTP cable are described by the graph shown in Figure 1-9, which shows average number of noise events per second exceeding some threshold [5]. This graph also shows how many packets (in percentage) one has to retransmit as a function of receiver margin assigned for the impulse noise. The distribution is truncated at some point (Vo).

Figure 1–9: Experimental Data, Impulse Noise



A packet loss rate allowable for FDDI link is $2*10^{-6}$ (based on packet loss rate allowable for FDDI network, consisting of 240 links, which is 0.05% [7]). That actually means that 1 noise event is allowed per

$$1/(2 * 10^{-6}) = 0.5 * 10^{6}$$

packets.

If the maximum FDDI packet consists of 45,000 bits, assume an average packet of $22.5 * 10^{3}$ bits, which occupies a time duration of $22.5 * 10^{3} * 8ns$ = 180 µs. One noise event must happen not more often than each of

$$0.5 * 10^{6} * 180 = 90 \ seconds$$

This equation also means that we allow 1/90 = 0.01 events (counts) per second.

According to Figure 1-9, 0.01 counts/sec corresponds to a threshold of -5 dBV or 560 mV.

Copper distributed data interface conditions can be different from those shown in Figure 1-9. The results shown in Figure 1-9 were obtained for 24 disturbing pairs suited on the same cable with the data carrying pair. Each interfering pair provided 20 telephone rings per hour, with 50 V pulses and 20 ns rise time and one pulse per ring.

Therefore, a data carrying pair gets strong cross talk from the close-located interfering pairs. The distance between the data carrying and interfering pairs can be as small as double insulation jacket thickness, which is 0.35 * 2 = 0.70 mm (for NT DIW cable).

If copper distributed data interface doesn't use transmission over voice carrying cable, then there will be a larger distance between two cables—data and voice carrying.

If a distance of two inches is assumed, then induced cross talk will be smaller by 25 mm * 2/0.7 mm = 71 times. Induced voltage is inversely proportional to distance between wires because field strength V/m decreases in proportion to the distance.

Therefore, instead of 560 mV, we obtain 560/71 = 7.9 mV, which is actually lower than the assumptions of 10 mV used for achievable distance calculations in Appendix B.

The fact that copper distributed data interface UTP bandwidth (30 MHz) is 1.5 times larger than 20 MHz will increase the threshold (margin) by the square root of (30/20) = 1.22 or 1.76 dB, assuming uniform spectrum distribution. This increase results in 7.9 * 1.22 = 9.64 mV. This number can be used as estimation of the external noise effects.

1.10 FREQUENCY RESPONSE

In general, the transmission attenuation constant (dB/m) for a coaxial cable [3] is

$$A = \frac{A_0 * \sqrt{f} \ [(1/a) + (1/b)]}{log(b/a)}$$

[3] where
a is the radius of the inner conductor
b is the radius of the outer conductor
A_0 is a constant.

According to the previous expression, if we can achieve a ratio of

$$b/a = 3.6$$

a minimum attenuation per unit length will result. The attenuation increases rapidly as a function of frequency and is a function of the square root of frequency, as shown in the previous equation.

According to the previous expression, the attenuation coefficient A(f) is proportional to square root of f [sqrt(f)] and can be rewritten as

$$A(f) = A_0 * \sqrt{f/f_o},$$

where A_0 is attenuation coefficient at frequency f_0 (dB/m). This model is commonly used also for balanced pairs. Consequently, the cable frequency response is

$$H(f) = 1/[A(f)]$$

The law for square root of f regulating the linear attenuation is equally valid at high and low frequencies for completely different reasons: the effect of the capacitance C (Figure 1-4) at low frequencies and the skin effect at high frequencies. The transition between the two regions is scarcely perceptible on certain lines but gives rise to a noticeable plateau on others.

1.11 EQUALIZATION

The frequency dependence of the cable's distributed resistance, R, causes more attenuation at high frequencies. An increase in channel bandwidth can be accomplished by employing a filter equalizer in the linear (analog) part of the channel. The filter transfer coefficient is designed to compensate for the frequency dependent resistive cable span loss. The coefficient has a magnitude that increases with frequency and nominally provides the equalization for an average length cable span. The intended result is for the cable-segment /equalizer to possess a flat frequency response characteristic over the system bandwidth.

An equalizer, placed in the transmitter, is emphasizing the high-frequency components of the signal so that when the channel selectively attenuates these, a clean signal is delivered to the receiver. In other words, the signal is predistorted to compensate for stronger cable attenuation of the high frequencies.

Equalization in transmitter—preshaping of the transmitted pulses is usually done with a transversal filter, which is actually a shift register. Successive bits of the incoming data are inserted into the shift register once every bit duration. The data bits are shifted through the shift register at a much faster rate. The parallel outputs are passed through multipliers with controllable gains and then fed to a summing device, which is followed by a low pass filter.

The clock that shifts the transmitted data stream has a higher frequency than the data signaling speed, so each pulse at the output is composed of a number of pulses (shifted in time) with different amplitudes, therefore creating a certain pulse shape.

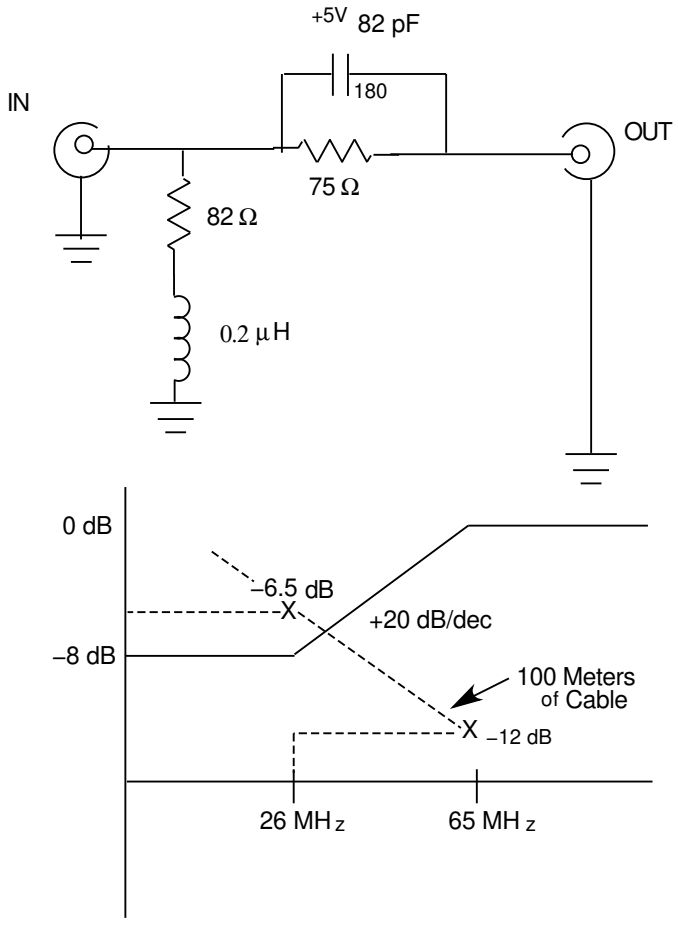
The spectrum of the staircase waveform is centered around the sampling frequency and its harmonics. By choosing a suitably large sampling frequency, the noise due to sampling can be separated from the spectrum of the desired pulses. Then a simple low pass filter can be used to smooth out the waveform.

Compared with receiver equalization, equalization in the transmitter has the advantage of not increasing the channel susceptibility to the external and thermal noise. In order to find the desired pulse shape, first we find the equalizer frequency response as the inverse of the cable response. Then we multiply the signal spectrum by the equalizer frequency response and this response would correspond to the desired transmitted spectrum. Finally, using Fourier transform, we obtain the waveshape to be transmitted.

In general, if the transmission length is limited by FCC requirements, it is beneficial to place an equalizer in the receiver to transmit as small a spectrum as possible. On the other hand, if external or thermal noise dominate, it is worthwhile to place most of the equalization in the transmitter to employ a receiver bandwidth as small as possible.

In coaxial and STP lines, due to protective shields, the external noise does not play a significant role and all the equalization functions can be placed in receiver. It can be a simple pole/zero network. The zero and pole are determined from an asymptotic approximation of the cables transfer function. An equalizer for a 100-meter link is shown in Figure 1-10.

Figure 1–10: Example of Equalizer Circuit



Note that an increase in high frequency gain results in an increase in thermal noise.

1.12 THERMAL NOISE CONTRIBUTION

The thermal noise contribution depends on equalizer frequency response.

The equalizer frequency response E(f) is the inverse of the cable frequency response

$$(E(f) \;=\; 1/H(f) \;= 10^{Ao[\sqrt{(f/fo)}]*l/20})$$

where l = distance.
If 4kT is the thermal noise spectral density, then the thermal noise power at the output of the equalizer is

$$P = 4\int_0^F kT\,[E(f)]^2\,df \;=\; 4kT\int_0^F 10^{0.1A_0*l\sqrt{f/f_0}}df$$

where F is the intended channel bandwidth.
Substitute $0.1 * Aol = a$ and $\sqrt{(f/fo)} = X$,

to obtain:

$$P = 4kT\int_0^{\sqrt{F/f_0}} 10^{aX}2f_0\;XdX = 8kTf_0\int_0^{\sqrt{F/f_0}} e^{2.3aX}XdX$$

which equals:

$$8kT\;\left\{\frac{e^{2.3a\sqrt{F/f_0}}\;[2.3a\sqrt{(F/f_0)-1}]\;+1}{(2.3a)^2}\right\}$$

This expression gives thermal noise power after the equalizer.

For example, if the bandwidth is F = 100 MHz and the attenuation of a 100-meter cable at f_0 = 62.5 MHz is 12 dB, or A_0 = 12/100 = 0.12.

At room temperature, T = 17°C or 290°K.

$$(kT = 4 * 10^{-21}(W/Hz))$$

The result is:

$$P \;=\; 8*4*10^{-21}\;*\;62.5*10^6*\frac{\left\{e^{2.3*0.1*0.12*100\sqrt{(100/62.5)}}\;[2.3*0.1*0.12*100\sqrt{(100/62.5)-1}]\right\}+1}{(2.3*0.1*0.12*100)^2}$$

$$= 21.76*10^{-12}(W)$$

If the equalizer is terminated on resistance R, RMS noise voltage is

$$RMS = \sqrt{PR}$$

For R = 50 ohm, RMS = $\sqrt{PR}$ = 33 µV.

To provide BER $\leq 2.5 * 10^{-10}$, the peak-to-peak signal V for antipodal signaling in white Gaussian noise must be at least

$$V \;= 12.6 * RMS = 0.42mV.$$

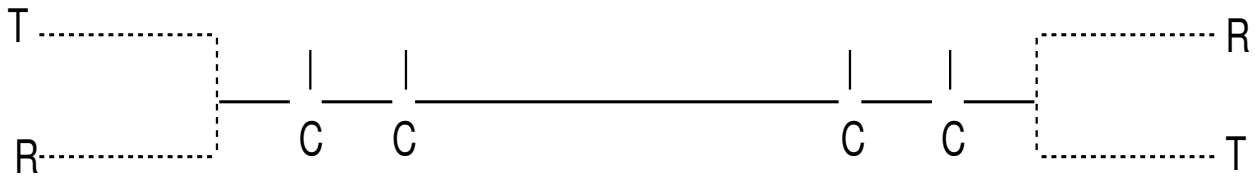
since

$$V_{pp} = 2\sqrt{2}V_{RMS}$$

1.13 TRANSMISSION LINK MODEL

Refer to Figure 1-11.

Figure 1–11: Transmission Line Model



where:

T =	Transmitter
R =	Receiver
S =	Transmitted signal level
K =	Line transfer coefficient
X =	Cross talk transfer coefficient
mS =	Total reflected signal level (sum of all the reflections)
BLW =	0.1S (baseline wonder of 4B5B); at the receiver we have 0.1KS
D =	Externally induced noise
N =	Peak-to-peak thermal noise after the equalizer. For probability $4 * 10^{-11}$ the peak-to-peak value of N = 13 RMS.

We have to provide the following conditions at the receiver:

$$KS - XS - D - 0.1KS - mS > N \qquad (1)$$

All parameters are peak-to-peak. Equation (1) can be rewritten in terms of minimum transfer coefficient as:

$$K = [N + D + S(X + m)]/0.9S \qquad (2)$$

Thermal noise (with moderate equalization) does not play a significant role in copper distributed data interface. For the above given example of the equalizer and assuming the minimum received signal level of 150 mV, we have to provide just 20 lg[(150 + 0.21)/150] = 0.01 dB additional margin for the thermal noise. Therefore expression (2) can be rewritten as:

$$K = [D + S(X + m)]/0.9S \qquad (3)$$

Now the achievable distance is:

$$L = k/d, \qquad (4)$$

where $k = 20\log(1/K)$ is allowable attenuation in dB and d is cable attenuation in dB/m.

1.14 TRANSMITTER RISE/FALL TIME

Due to presence of both cross talk and ISI (inter-symbol interference) because of insufficient cable bandwidth, there is an optimum in transmitter rise/ fall time. The reasoning is that when the rise/fall time is getting lower, cross talk is increased (because higher frequencies are now present), but ISI is decreased, therefore providing a largest eye opening at some rise/fall time value. Because the cross talk and ISI may affect the eye diagram differently in vertical and horizontal planes, some kind of optimum eye opening (after the equalizer) can be determined, and the numbers for the optimum rise/fall time can be specified.

This approach is especially acute in the coaxial version, because, in an STP link, the cross talk is very low and in UTP, the limitations on emission limit the transmitted bandwidth (rise time).

1.15 DCD, BASELINE FLUCTUATIONS, AND HYSTERESIS IN THE SIGNAL DETECT CIRCUITRY

If 4B5B data contain frequencies below the receiver low frequency cutoff, there will be baseline fluctuations of ±10%, that decrease the horizontal eye opening. The effect of these fluctuations can be attributed to DDJ or to DCD, measured at 40 or 60%. Therefore, the largest DCD between 40 and 60% levels, is accountable for eye opening reduction in time domain due to both factors: DCD and baseline fluctuations.

If compared with classical FDDI, which accounts for baseline wonder and thermal type of noise, copper distributed data interface has additional impairments: cross talk, reflections, and external noise. These impairments require an additional margin for the hysteresis in the signal detect circuitry.

1.16 COAXIAL CABLES

Coaxial cable impedance is dependent on the cross-sectional geometry of the cable and its dielectric properties. Because both the inductance and capacitance depend on the ratio of the radius of the outer conductor to the inner conductor, it is possible to increase both the inner conductor and the outer conductor radii so as to keep the characteristic impedance constant but still reduce the resistance R.

The resistance R is dominated by losses on the inner conductor and this resistance value is inversely proportional to the inner conductor radius. Selecting a larger inner conductor diameter will reduce the ohmic losses at all frequencies. The dc resistance of the center conductor is inversely proportional to its cross-sectional area and a larger conductor gives lower dc losses as well.

In a given cable, the choice of materials for the insulating dielectric, center conductor, outer shield conductor, and outer insulating jacket depends on the application for that cable. The dielectric material providing insulation between the center and the shield conductors can be polyethylene, Teflon, plastic foam, or air. Center conductors can be copper or aluminum. The outer shield conductor can be made of copper or aluminum braid, foil, or tubing. The outer insulating jacket, when required, is usually made of polyvinyl chloride (PVC).

The coaxial cable as shown in Figure 1-3 has a solid center conductor surrounded by a dielectric that is in turn enclosed by a conducting shield. The currents run on the outside surface of the inner conductor and the inside surface of the outer conductor. The dielectric functions as an insulator.

The characteristic impedance of cables, intended for FDDI applications, is Z_0 = 50 ohms.

1.17 SHIELDED BALANCED PAIRS

Refer to Figures 1-12 through 1-15. The following cables are considered for copper distributed data interface.

- **IBM Type 1**—22 American Wire Gauge (AWG) (solid) bare copper; 4 polyethylene-insulated conductors in 2-pair parallel; each pair foil shielded; overall braid shield of tinned copper; PVC jacket with nylon ripcord; attenuation at 16 MHz is 4.5 dB/100 m; cross talk is -40 dB at 20 MHz.
- **IBM Type 2**—22 AWG (solid) bare copper; 12 polyethylene-insulated conductors in six pairs; Two foamed data grade media pairs in parallel, each foil shielded (optional); an overall tinned copper braid shield; four solid voice grade media pairs, cabled around the data grade core in parallel; PVC jacket with nylon ripcord; attenuation at 16 MHz is 4.5 dB/100 m; cross talk is -40 dB at 20 MHz.
- **IBM Type 6**—26 AWG bare copper; 4 polyethylene-insulated conductors in two twisted pairs laid parallel; each pair foil shielded; overall braid shield of tinned copper; striated PVC jacket; attenuation at 16 MHz is 6.6 dB/100 m; cross talk at 20 MHz is 34 dB.
- **IBM Type 9**—26 AWG solid bare copper two twisted pair in parallel (0.018 inch OD), overall foil and polyester tape, 65% tinned copper braid overall and a solef jacket with nylon ripcord. Therefore, it does not have shields for individual pairs.

Frequency response, group delay, and cross talk attenuation are shown in Figures 1-16 and 1-17.

Figure 1–12: Cable Types

Type 1 specification number 4716748
Type 1 Plenum specification number 4716749
Braided cable shield around two twisted pairs of number 22 AWG conductors for data communication.

Type 1 Outdoor specification number 4716734
Corrugated metallic cable shield around two twisted pairs of number 22 AWG conductors for data communication Type 1 outdoor cable is suitable for aerial installation or placement in conduit underground.

Type 2 specification number 4716739
Type 2 Plenum specification number 4716738
Same as type 1 cable with the addition of four twisted pairs of number 22 AWG telephone conductors.

Type 6 specification number 4716743
Two twisted pairs of number 26 AWG stranded conductors for data communication. Type 6 cable is for use only as patch cables or jumper cable.

Figure 1–13: Type 1 Outdoor

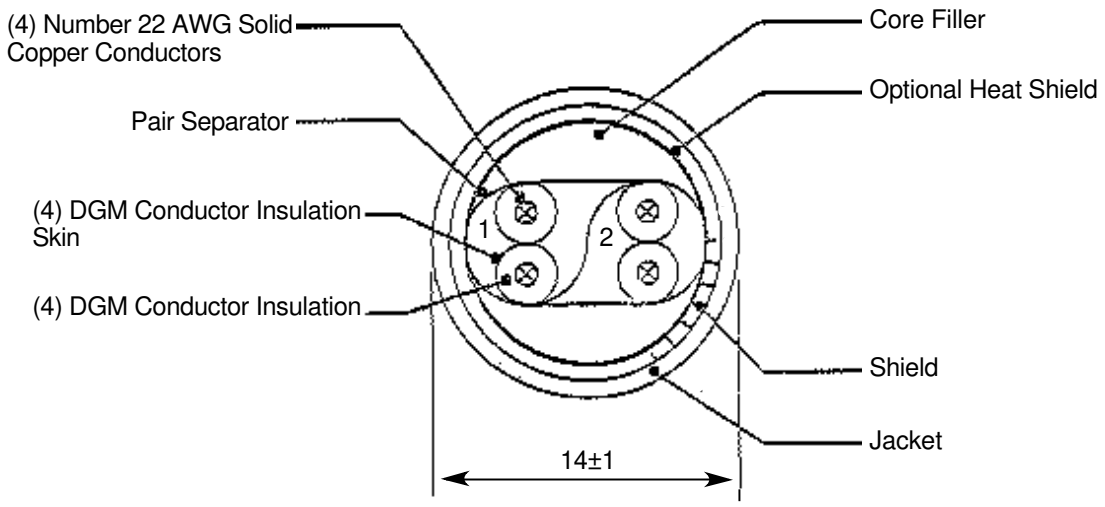


Figure 1–14: Type 1 Plenum, Non-Plenum, Riser

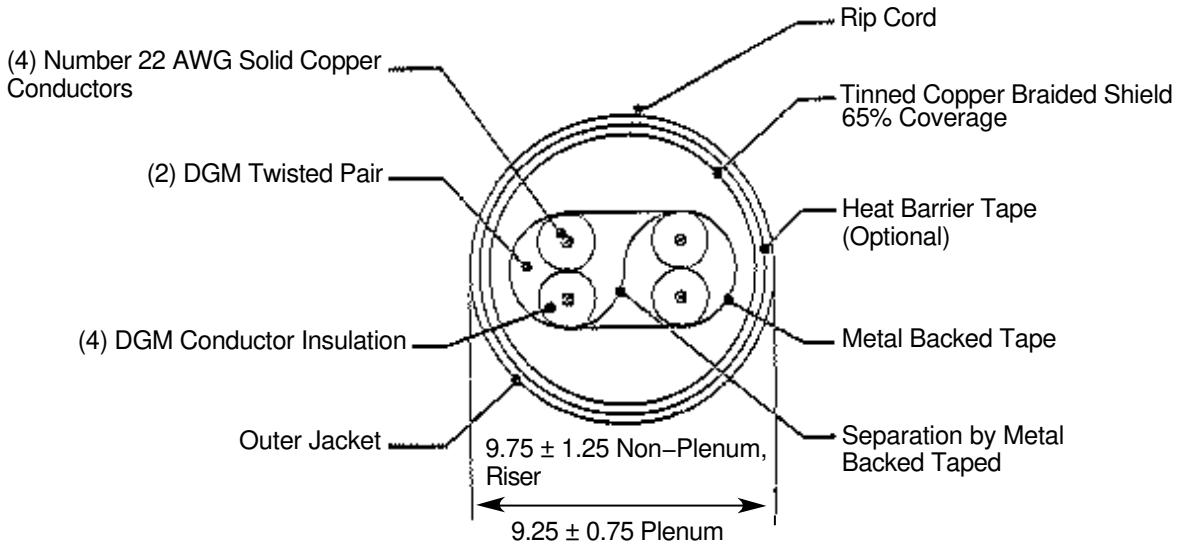

(4) Number 22 AWG Solid Copper Conductors
(2) DGM Twisted Pair
(4) DGM Conductor Insulation
Outer Jacket
Rip Cord
Tinned Copper Braided Shield 65% Coverage
Heat Barrier Tape (Optional)
Metal Backed Tape
Separation by Metal Backed Taped
9.75 ± 1.25 Non–Plenum, Riser
9.25 ± 0.75 Plenum

Figure 1–15: Type 2 Plenum, Non-Plenum

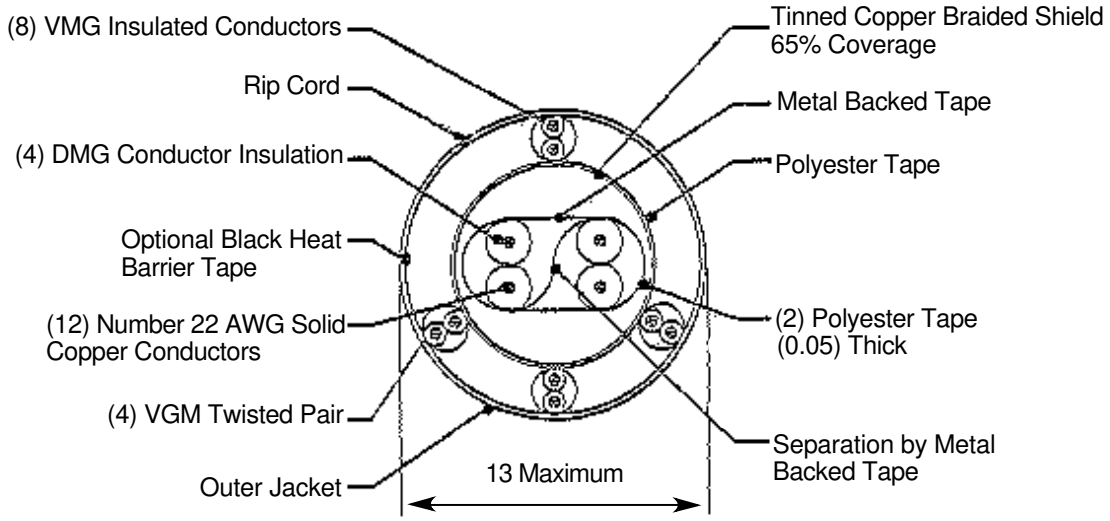


Figure 1–16: Type 6 Office

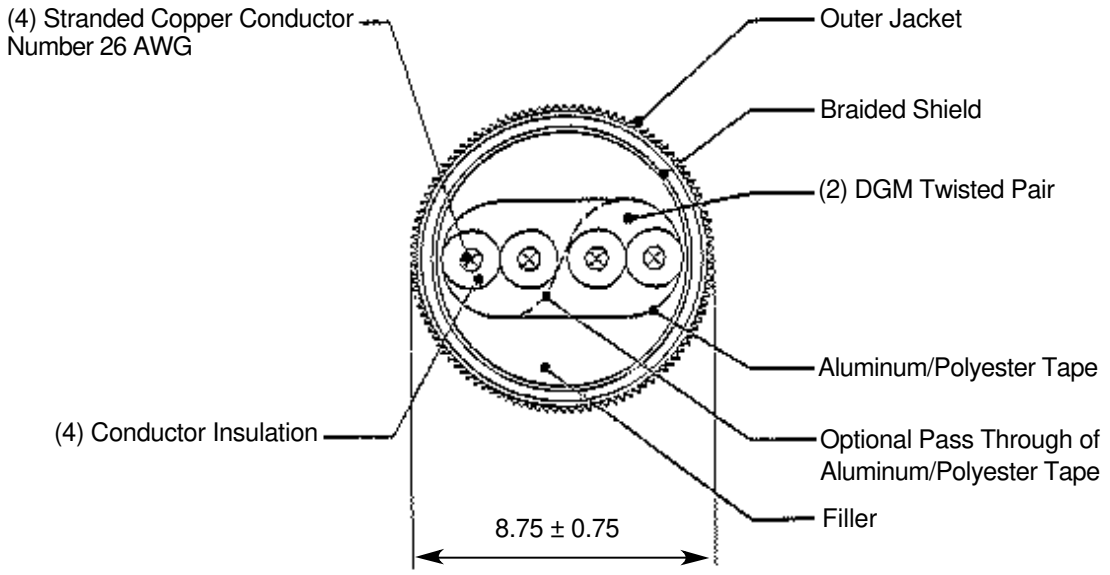


Figure 1–17: Attenuation and Group Delay, 112 Meters—Type 6

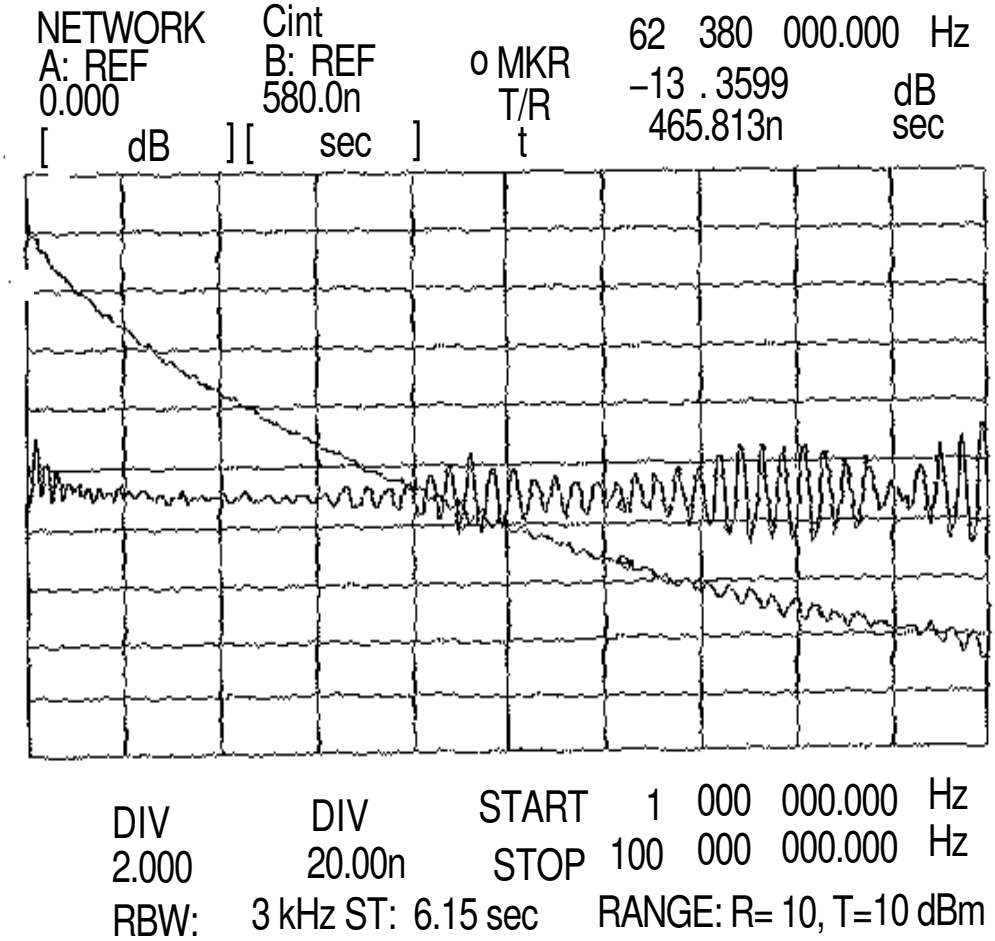


Figure 1–18: Near End Cross Talk—Type 6

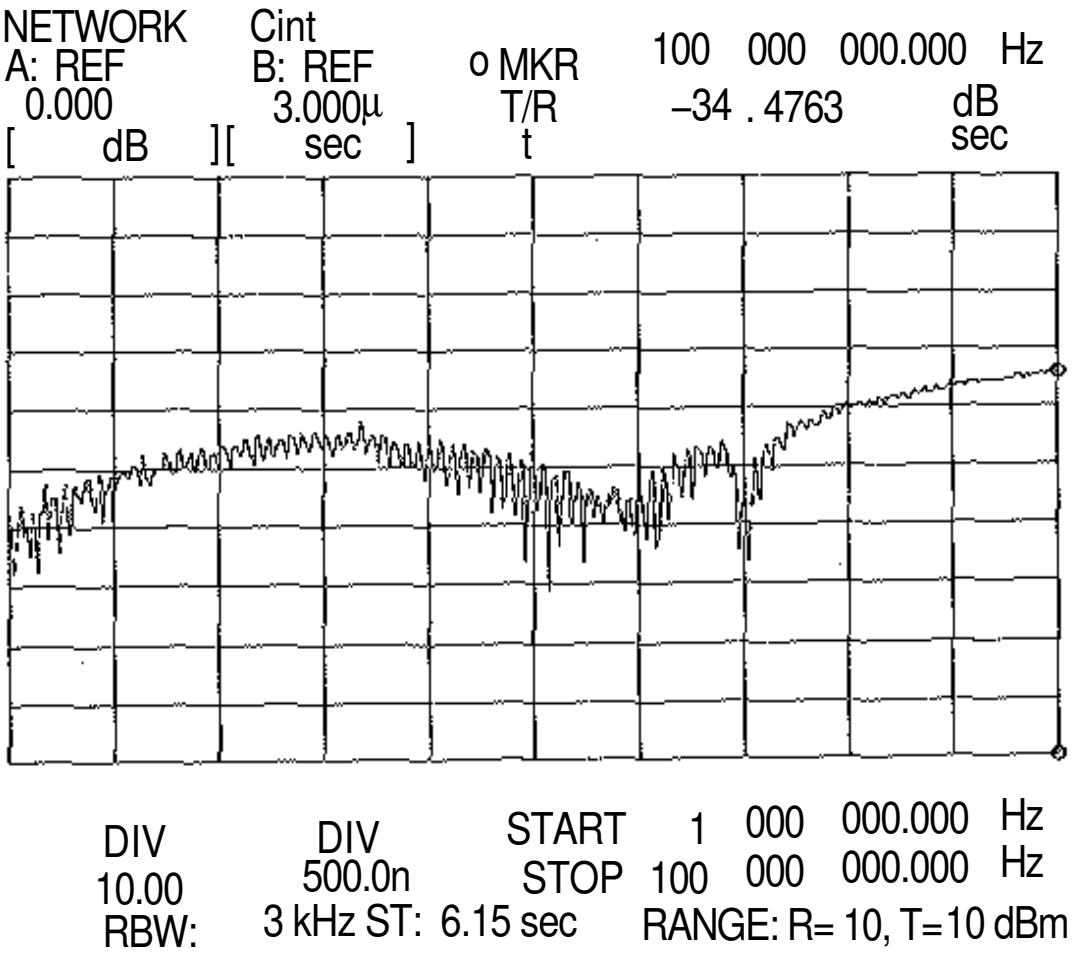


Figure 1–19: Attenuation and Group Delay, 100 Meters—Type 1

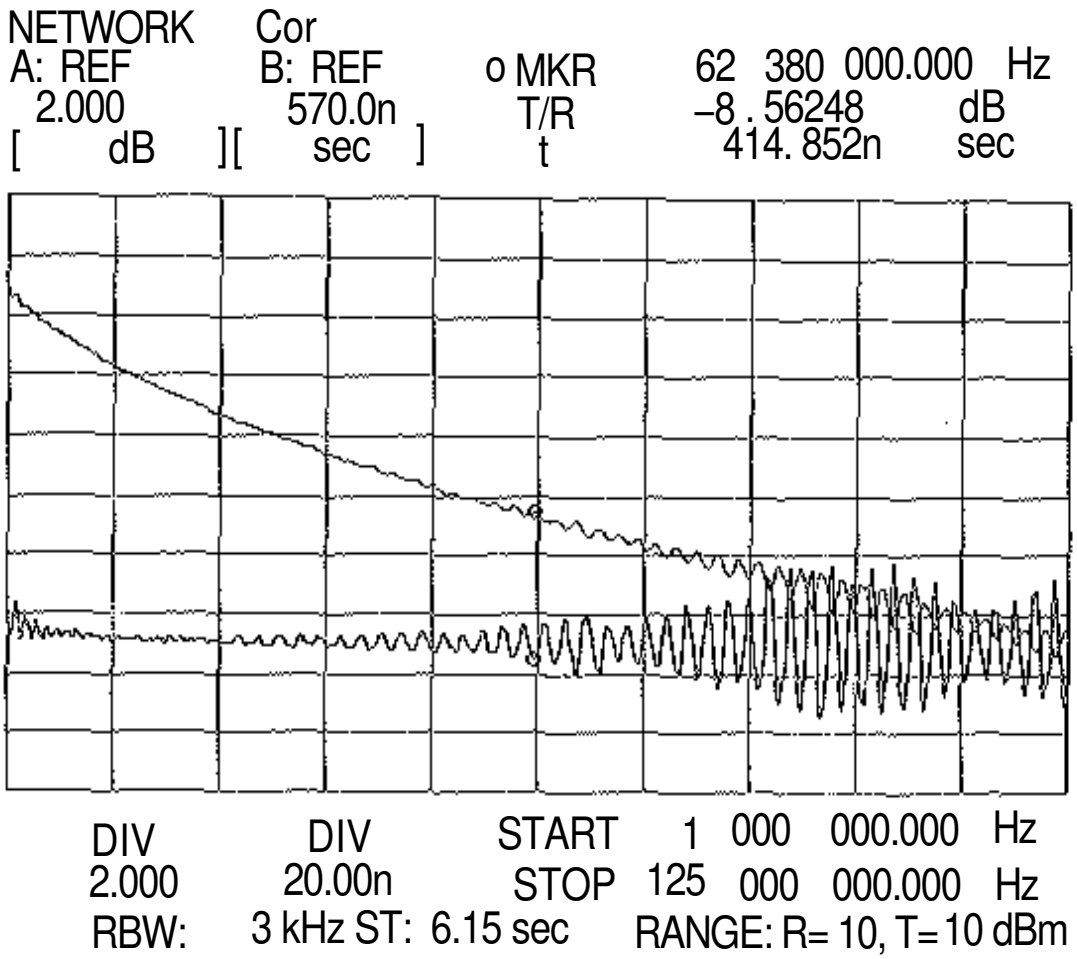
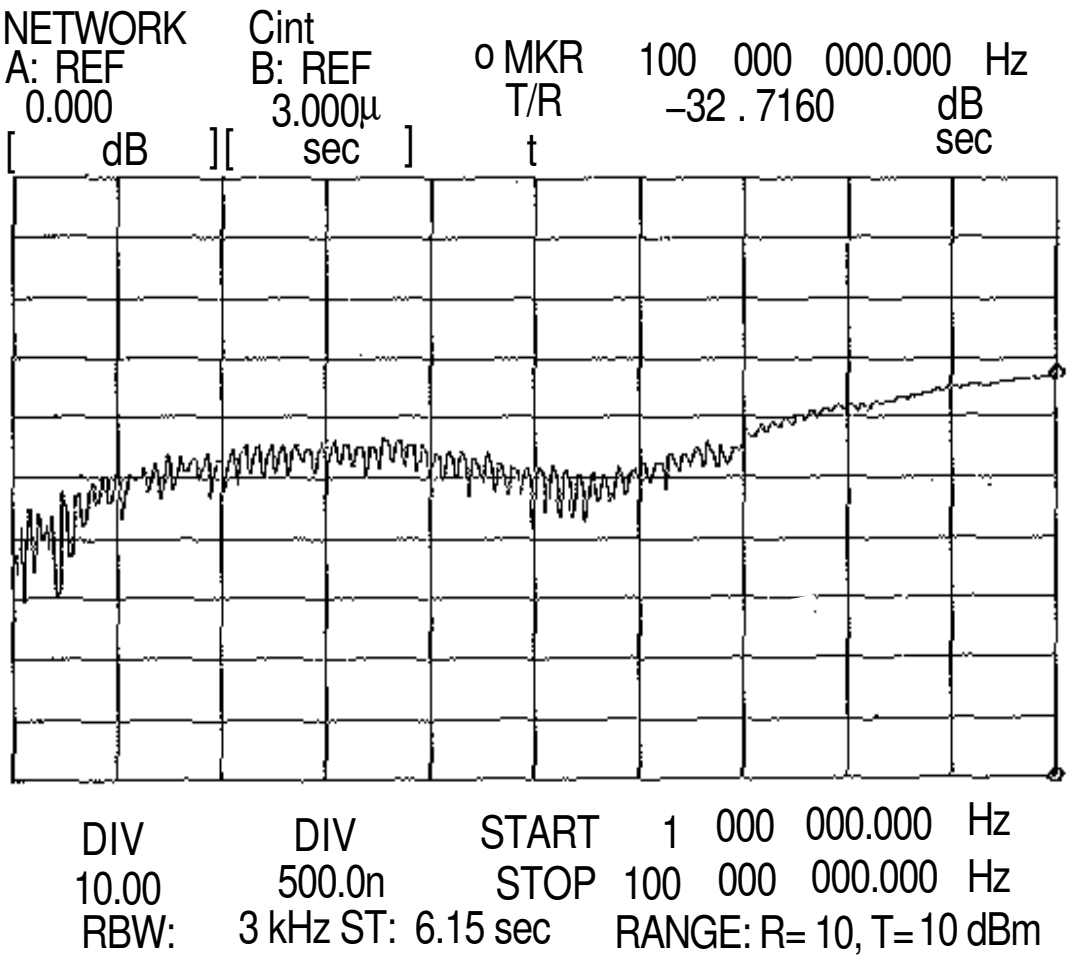


Figure 1–20: Near End Cross Talk—Type 1



1.18 UNSHIELDED TWISTED PAIR (UTP)

A key argument in favor of UTP is that it is a ubiquitous medium.

The most widespread is old analog phone wiring—24 AWG (0.5 mm). Also, larger 22 AWG and smaller 26 AWG cables are in use.

The required operating distances must not be less than 50 m to provide connections to at least 95% of networked workstations (they reside within 50 m of the wiring closet [8]).

High on the list of potential concerns is the quality of the wire that varies widely from facility to facility and affects cross talk, electromagnetic interference (EMI), and cable attenuation.

EMI is a concern at two levels. First, the wire itself is clearly affected by EMI emitted from external sources, for example, a fluorescent light fixture; EMI will affect the signals being transmitted on a wire. Second, UTP wire itself radiates electromagnetic energy. The amount of allowable radiation must not exceed the FCC requirements.

UTP is actually a pair of solid wires or wire strands twisted around each other and surrounded by unshielded insulation jacket. Some versions contains overall shield.

The twists in the wire create coupling, which cancels out opposing electromagnetic fields. The greater the number of times the wire twists around itself in a given distance, its "twist ratio", the tighter the coupling.

When a wire is loosely twisted, there is less coupling and greater chance of cross talk leaking into adjacent wires [1]. In general, the relationship between near-end cross talk and frequency is not linear and not monothonical, therefore complicating cross talk compensation.

Unlike NEXT, attenuation (in dB) does have a linear relationship with frequency. As frequency increases, the level of attenuation increases.

Traditionally, UTP cables were specified for voice frequency range. So the old installed cable specifications are not adequate for FDDI related frequencies, and UTP cables exhibit unpredictable behavior from cable to cable for FDDI frequency range.

[1] Sometimes the cross talk is created by a negligence. At the punch block for example, installers often leave several inches of untwisted wire where the shielding has been stripped from the cable to make a connection to the block. Another example is the use of patch cord (silver satin), which typically connects a telephone to the wall jack. An incorrect logical pairing of wires in the connectors can create a split pair, where wires run in parallel, eliminating the necessary twist ratio.

Some specific examples are described in the following paragraphs:

Wire Diameter

Manufacturers used to specify the wire nominal diameter and minimum, but not maximum or tolerances. For instance, 24 gauge means nominal 0.5 mm and minimum (NT) 0.35 mm. corresponding to the specified minimum conductor resistance. The skin effect will result in different signal attenuation, cross talk and radiation properties at high frequencies for two wires with different diameters and the pair actually will be unbalanced. In addition, the lumped resistance is different, as well as lumped inductance.

Distance Between Wires

Manufacturers used to specify nominal and minimum wire jacket thickness (for 4-pair NT cable nominal thickness is 0.64 mm and minimum 0.35 mm and no limit on maximum). therefore the distance between wires can fluctuate in large range.

Emission power (in dBm) from "balanced" line is directly proportional to distance between wires squared and frequency squared. If the distance slightly changes over the cable length or from cable to cable, we will get different emission properties.

Therefore, the way to provide reliable transmission is to get extensive margin to pass a cable with unknown characteristics.

UTP cables are presently classified into five categories with regard to specified frequency range [9].

Category 5 includes cables, for which characteristics are specified up to 100 MHz.
Category 4—up to 20 MHz
Category 3—up to 16 MHz
Category 1 and 2—cables for voice and low-speed data transmission rates.

The following parameters of the cables consisting of four unshielded twisted pairs of 24 AWG thermoplastic insulated conductors and enclosed in thermoplastic jacket are specified:

- Near end cross talk, in the form of tables or graphs as a function of frequency
- Mutual inductance of any pair at 1 kHz at 20°C shall not exceed 20 nH for category 3 and 17 nH - for categories 4 and 5, if measured over 305 m.
- Characteristic impedance is 100 ±15% (mostly due to structural variations). The structural return loss of the line terminated by 100 ohm shall be greater than 14 dB.
- The maximum attenuation at the highest specified frequency shall be less than 40 dB for category 3, 31 dB for category 4, and 67 dB for category 5, if measured for 305 m at 20°C.

  Temperature dependence is described by factor of 0.3% per degree Celsius for categories 4 and 5.
- Near end cross talk loss shall exceed 23 dB, 36 dB, and 32 dB respectively, for the highest specified frequency for categories 3, 4, and 5.

The frequency dependence is determined by the expression:

```
NEXT(F) > NEXT(Fo) - 15 log (F/Fo).
```

Signal attenuation and cross talk as functions of frequency are shown in Figure 1-21 and 1-22.

Figure 1–21: Attenuation Measurements

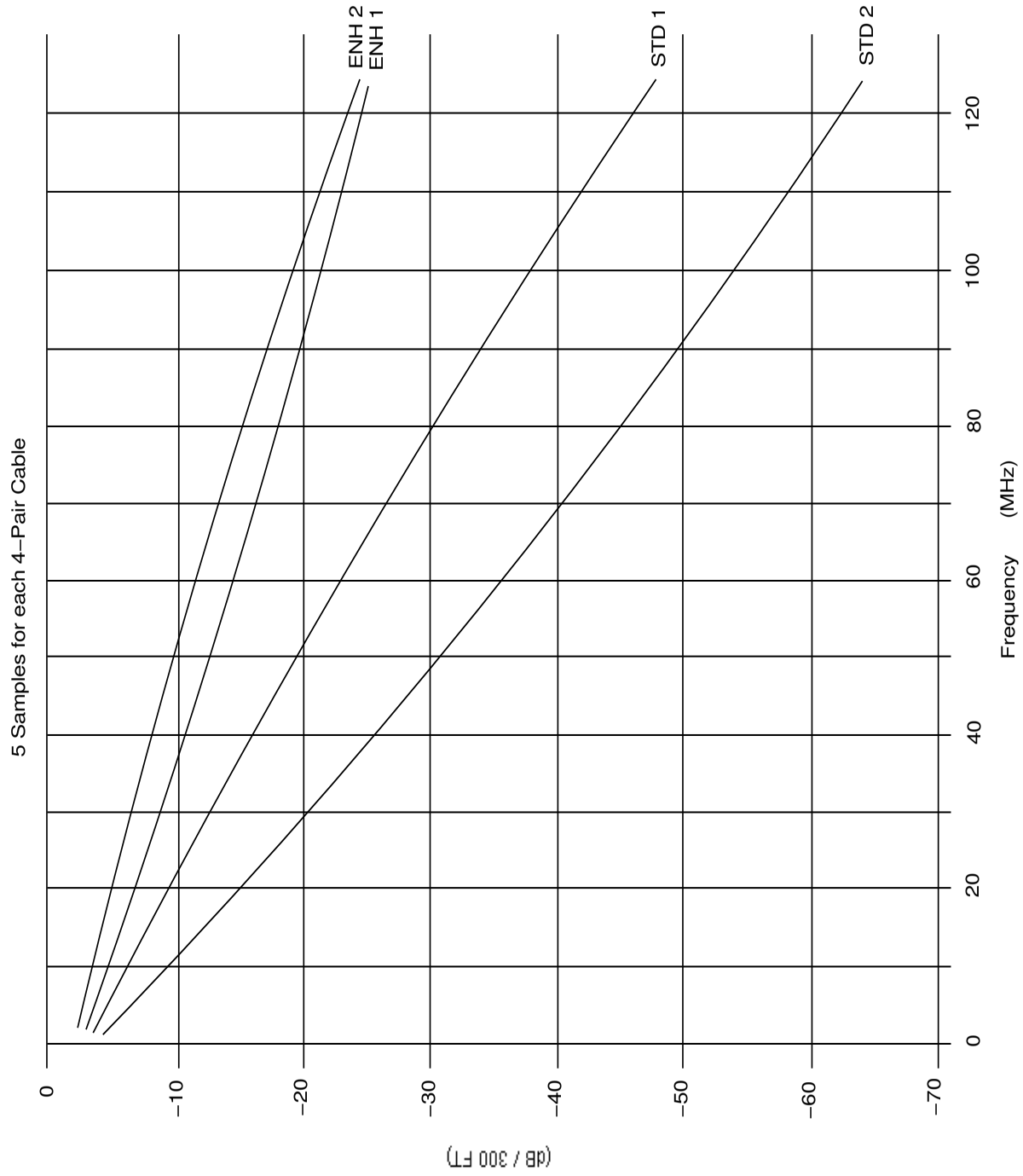
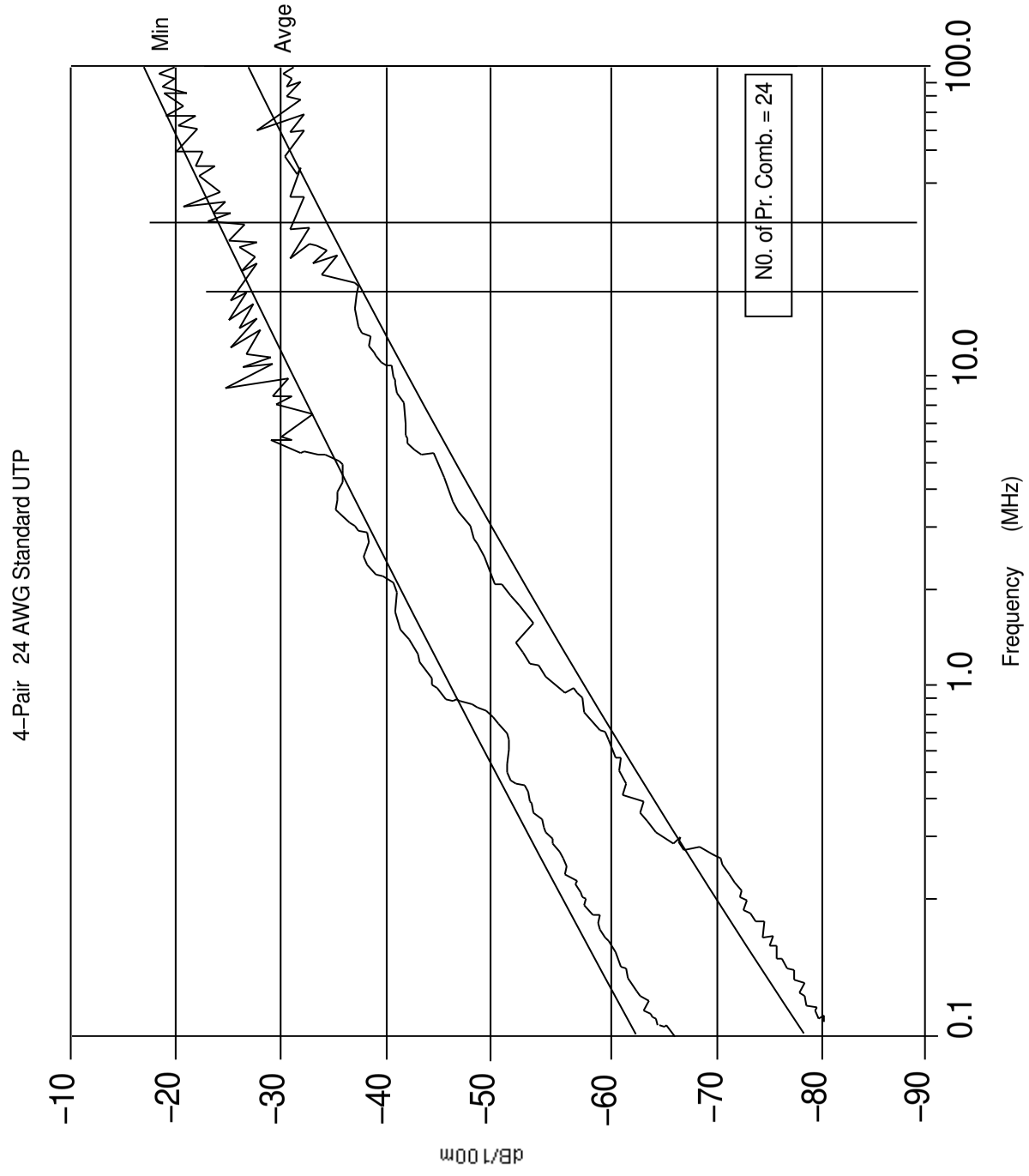


Figure 1–22: Near End Cross Talk Measurements



CHAPTER 2

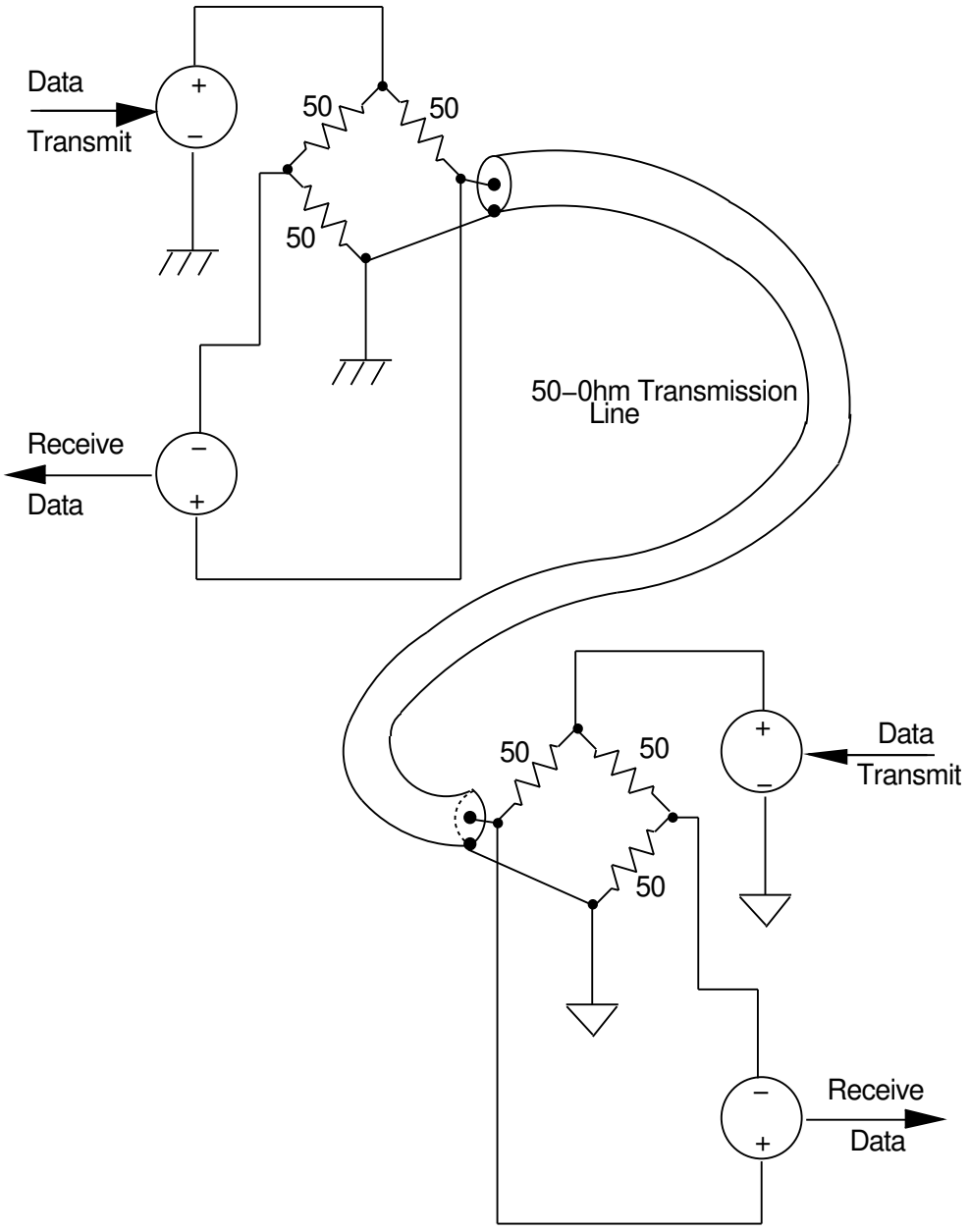
COAXIAL LINK

2.1 BIDIRECTIONAL TRANSMISSION

To establish bidirectional and simultaneous data communication over a single coaxial wire, a hybrid bridge configuration is used [10, 11, 12]. A 50-ohm network is configured so that the locally generated transmit signal is subtracted out of the signal appearing on the wire. Refer to Figure 2-1. Ideally done, the resulting received signal is that of the distant transmitter at the other end of the link. Because exact cancellation depends upon the impedance of the cable, reflections from impedance junctions, and the attenuation of the cable, the length will be limited, depending upon the configuration.

The worst-case configuration occurs when a round trip from transmitter to a connector is equal to an integer number of the bit duration T. Because the signal delay is about 4 or 5 ns per meter, the worst case occurs if distances to the connectors are about to be multiples of 1 m. Because the link may have up to three closely located connectors, the transmitted wave can be corrupted by three reflected waves — all three might have opposite phase, if compared with the incident wave. Therefore, in the worst case, two reflected waves will be subtracted from the incident wave. This subtraction happens at some bits and may not happen in others because the reflected wave differs from the incident wave. Assuming equal probabilities for 1s and 0s, the probability of the worst case is $(0.5)^4 = 0.0625$. That means that, on the average, each 16th pulse-bit will face this worst-case reflection.

Figure 2–1: Bidirectional Transmission



2.2 REQUIREMENTS TO RETURN LOSS OF JOINTS IN COAXIAL LINE

Requirements to a connector return loss (P) depends on the distance from the connector to the closest transmitter. Note that the transmitters are on both sides of the line, as well as the receivers.

In equation form:

$$P(dB) = G(dB) - k(dB)$$

where k = cable attenuation and G = minimum allowable return loss for a connector.

Because there are line receivers on both sides of the line, k is counted as an attenuation of round trip from the closest transmitter to the joint.

For example, if cable attenuation is 0.1 dB/m, assuming a connector is 80 m from one transmitter and 20 m from the other (the link's length is 100 m) and the minimum allowable return loss is 25 dB, then the required connector return loss is more than or equal to

$$25 - (20m * 2 * 0.1dB/m) = 25 - 4 = 21dB$$

If there are n closely located connectors, the required minimum connector return loss is

$$P(dB) = G(dB) - k(dB) + 20\ lg\ n$$

2.3 ACHIEVABLE DISTANCE

Because the externally induced noise is negligible, expression (3) of subhead 1.13 modifies to

$$K = S(X + m)/0.9S$$

For example, if we assume cable attenuation of 12 dB/100 m, S = 1 V, cross talk (through the bridge) attenuation 20 dB (10 times) and the total return loss is 20 dB, X = 0.1 and m = 0.1)), we obtain:

$$K = 1 * 0.2/0.9 = 0.22; \quad 1/K = 4.5(13.1dB)$$

The allowable distance is:

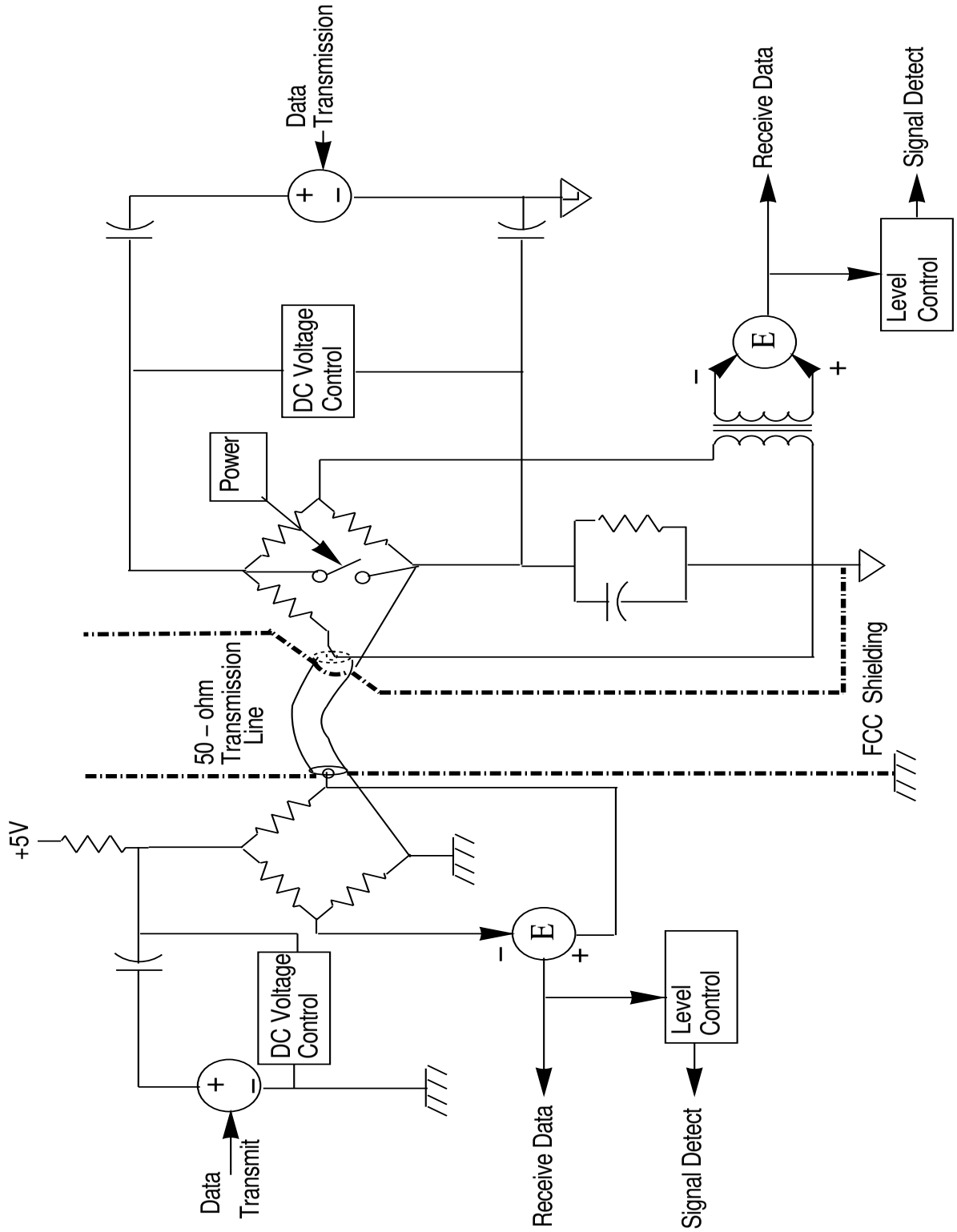
$$13.1/12 = 109m$$

Appendix A contains data on distance, limited by reflections, for different configurations of the cable set.

2.4 CHANNEL STRUCTURE

A block diagram of a point-to-point coaxial transceiver link is shown in Figure 2-2. The link consists of two bidirectional coaxial transceivers, a length of coaxial media, and two media interface connectors (MICs).

Figure 2–2: Point-to-Point Coaxial Transceiver Link Block Diagram



2.5 DATA DEPENDENT JITTER (DDJ)

DDJ plays a significant role in classical FDDI because it has a large additional component due to the cross talk. If we assume signal rise/fall time = 3.5 ns and the magnitude of cross talk is 6 dB below the signal level, then the cross talk jitter is 3.5 ns x 0.5 = 1.75 ns.

2.6 LINK LOSS DETECTION

The function of the link detection is to prohibit Signal Detect assertion due to the reception of ones own reflected signal if the coaxial line is improperly terminated. See Figure 2-3. The dc circuit model is shown in Figure 2-4.

Figure 2–3: Coaxial Line

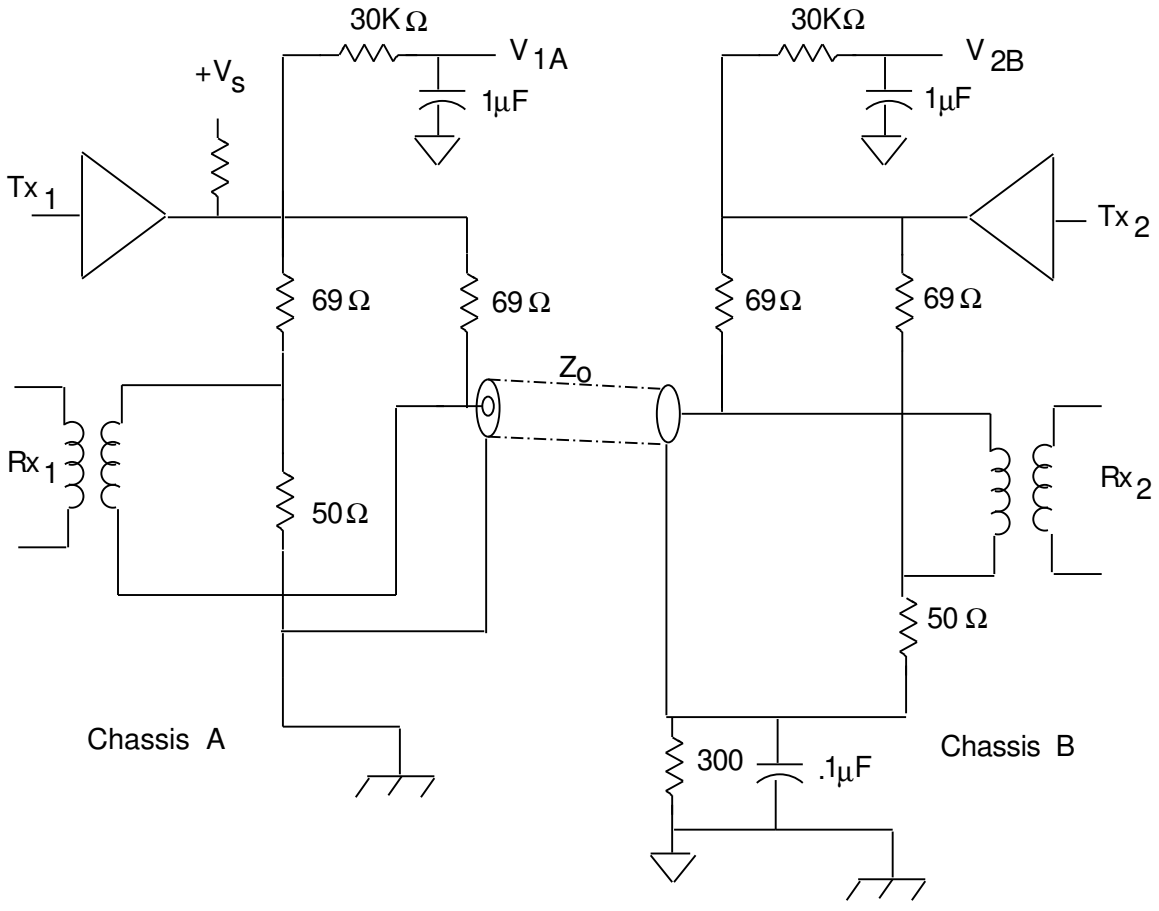
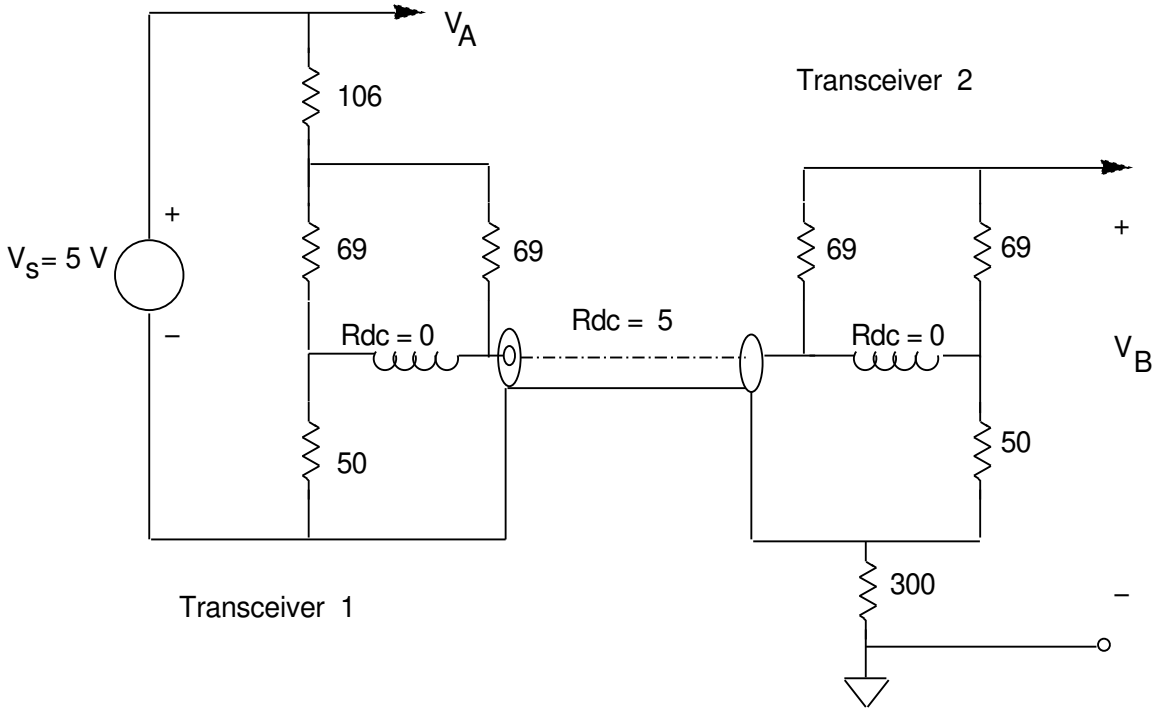


Figure 2–4: Link Detect DC Circuit Model



CHAPTER 3

SHIELDED TWISTED PAIR (STP) LINK

3.1 INTRODUCTION

The STP link is intended to connect FDDI concentrators to other FDDI concentrators or to FDDI stations, using a 150-ohm shielded twisted pair shown in Figure 1-11. The cable is Type 1 and 2 of IEEE 802.5 Token Ring networks. Type 6 may be used for short patch cables, as shown in Figures 1-12 and 1-13.

Both attenuation and cross talk as a function of frequency, are shown in Figure 1-14 through 1-17.

3.2 LINK BLOCK DIAGRAM

The functional organization of the STP link is shown in Figures 3-1 and 3-2. The link consists of two transceivers, including the link detection circuit, media interface connectors (MIC) and a cable plant [11, 13, 14].

Two port types, labeled *M-Port* and *S-Port*, are defined. The shield is tied directly to the chassis at both the M-Port and S-Port through a low-impedance path.

Figure 3–1: Shielded Twisted Pair FDDI

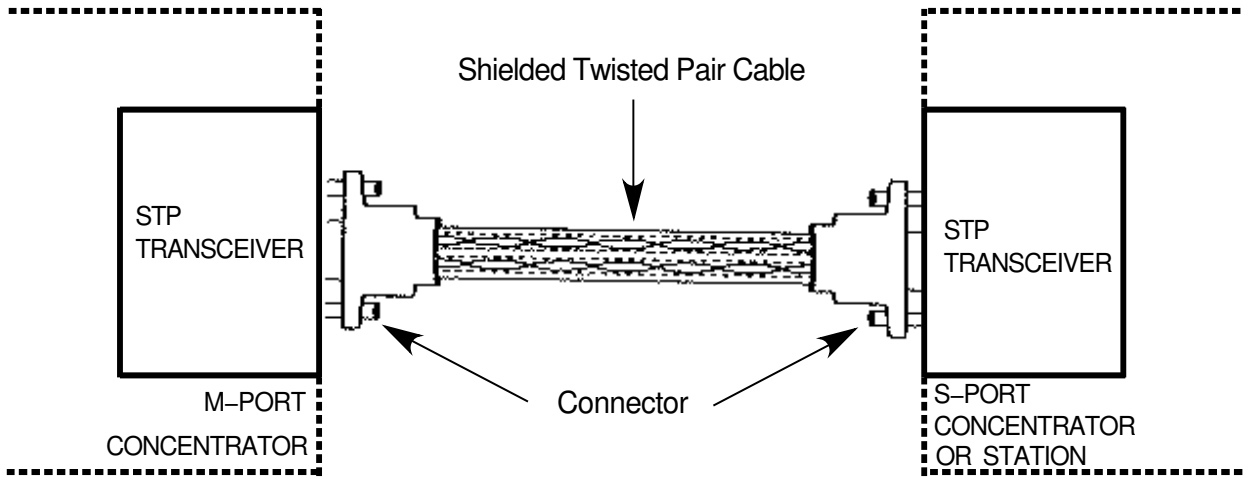


Figure 3–2: STP Link Functional Diagram

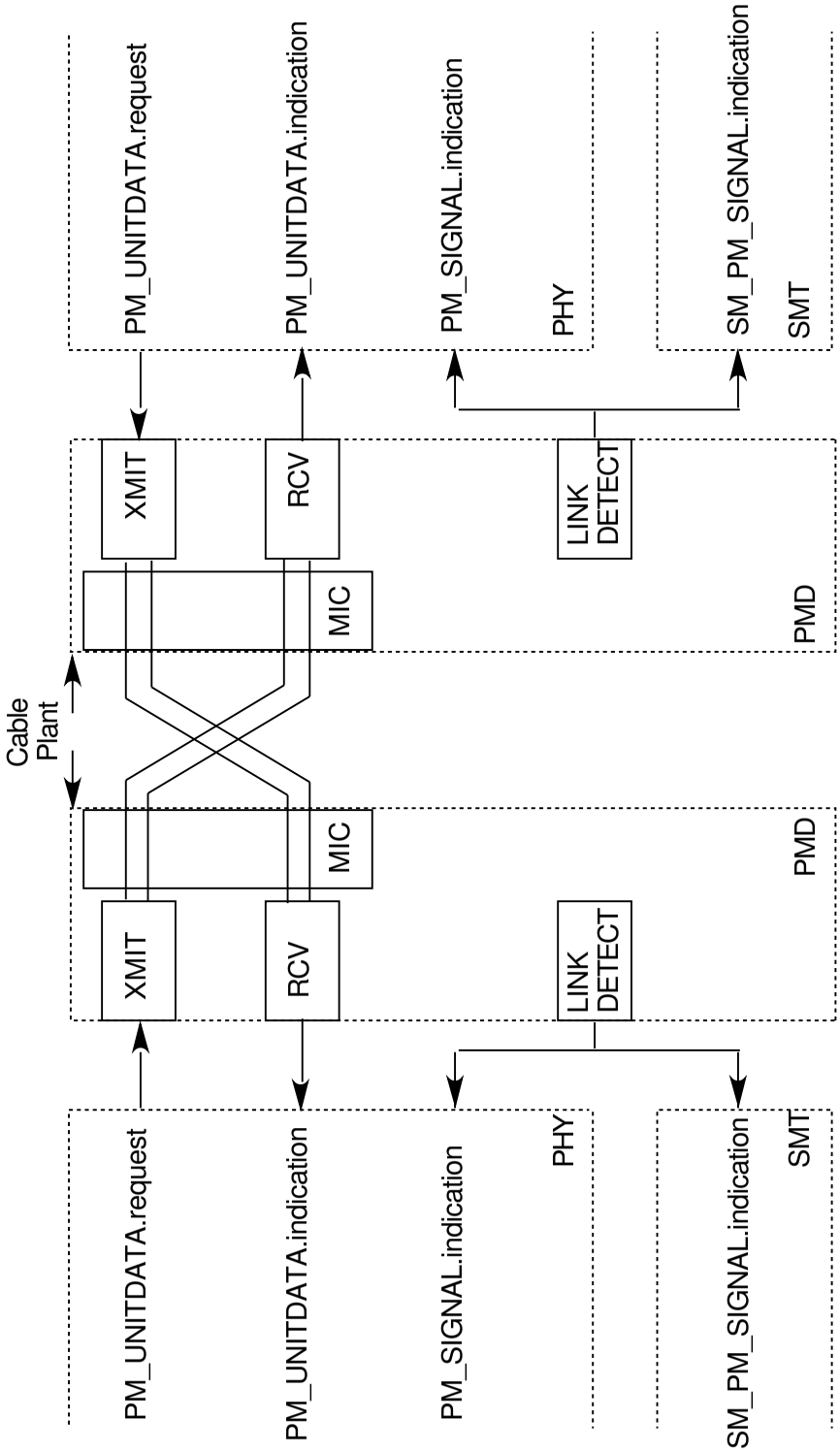
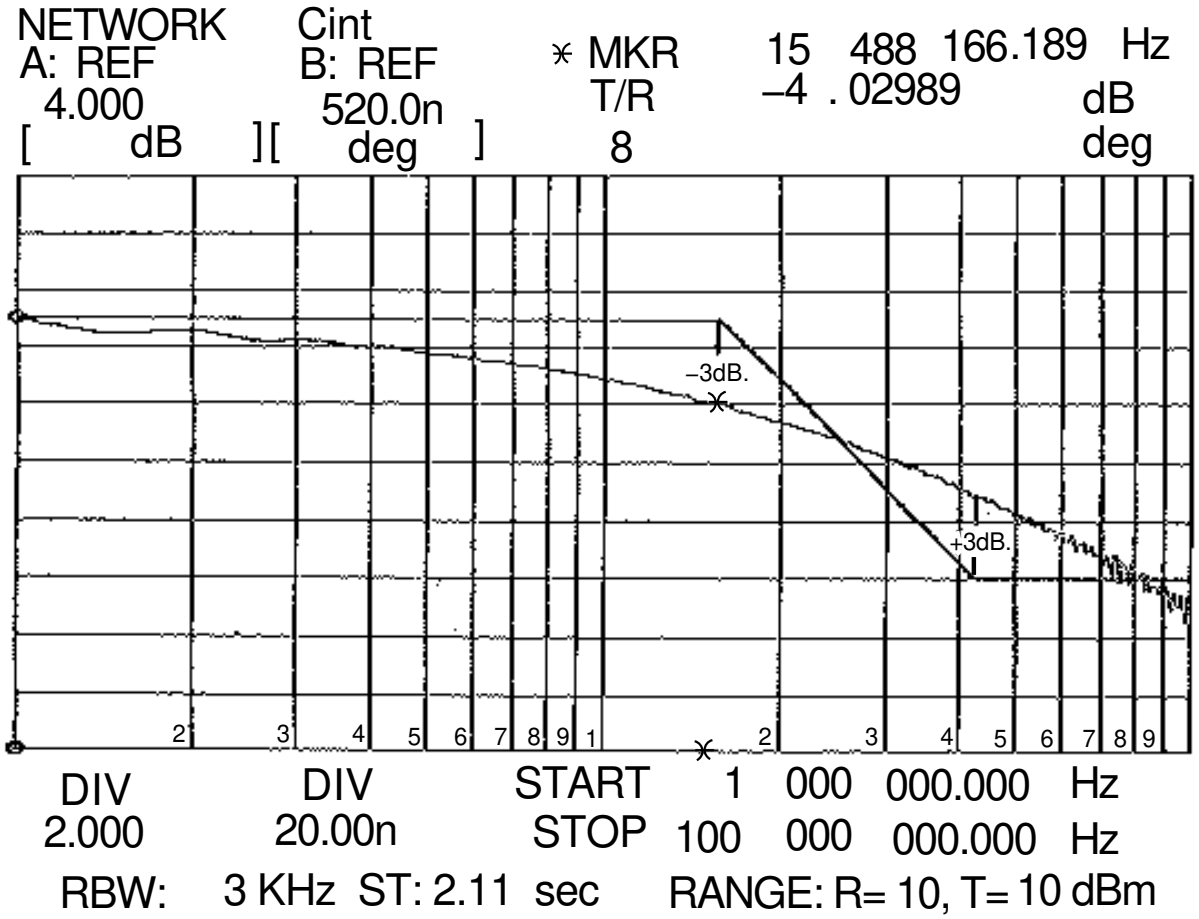


Figure 3–3: Asymptotic Zero/Pole Approximation to Cable

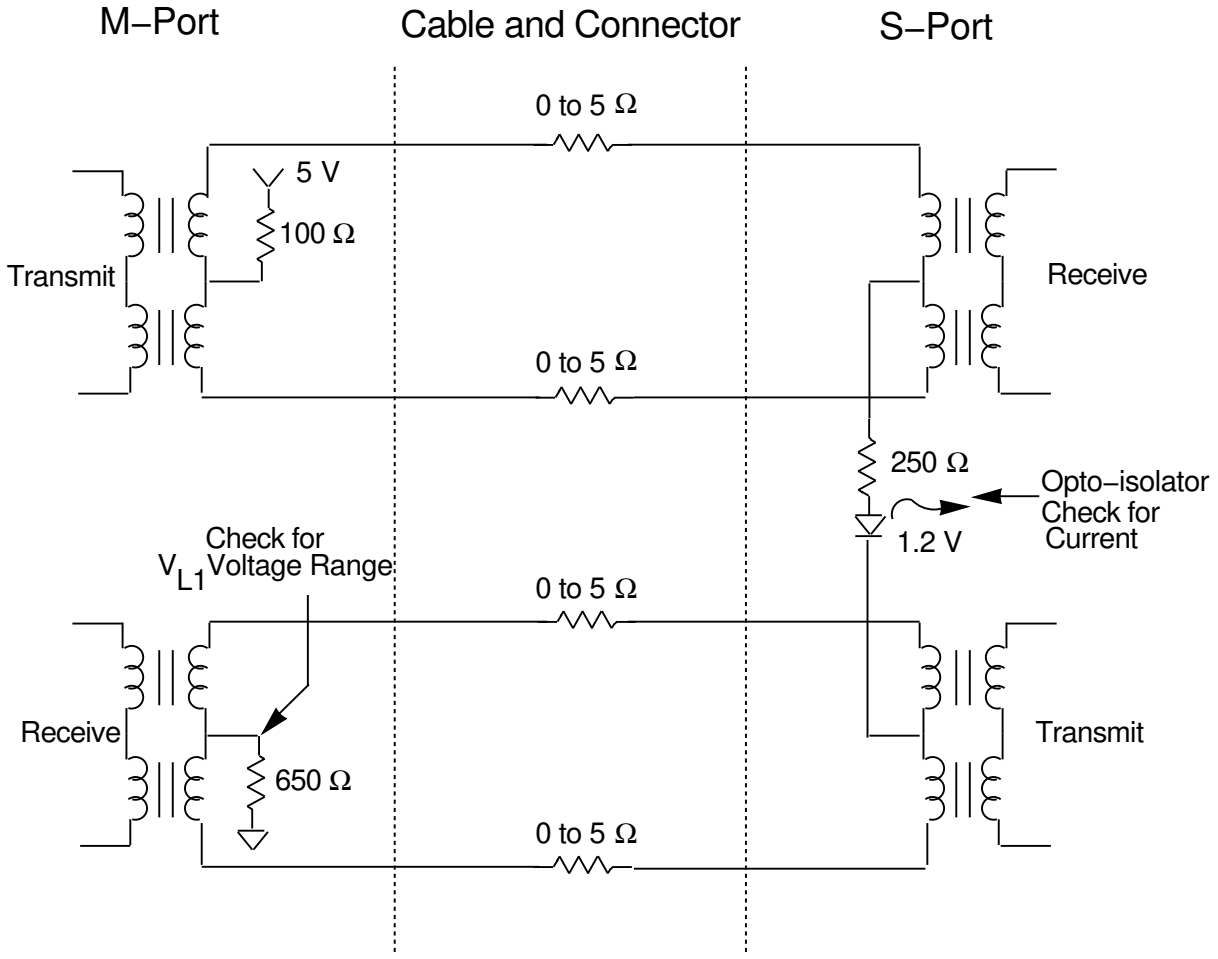


3.3 LINK DETECTION CIRCUITRY

The function of the link detection circuit is to ensure that Signal Detect assertion is caused by the reception of a link partner's signal. The other function of the link detection circuit is to indicate sufficient received signal is present to satisfy a minimum bit rate. Practically, the link detection circuit function is the logical AND of the signal detect function, which determines a sufficient received signal is present, and the cable detect function, which determines the ports are connected properly. The signal detect (status) parameter of the PHY is set by the link detect function.

The cable detect circuit is shown in Figure 3-4.

Figure 3–4: Cable Detect Implementation Example



Transceiver M supplies a phantom current of approximately 4 mA through transceiver S, and then back to transceiver M. A loopback condition is determined by monitoring the signal V_r for voltage values of approximately 4.3 V. Normal link operation is indicated when V_r is about 2.5 V and the opto-isolator is being driven by about 4 mA.

Contrary to a fiber optic link, the cross talk and reflections play a significant role in the STP link and they actually determine the thresholds of the signal detect circuitry.

3.4 ACHIEVABLE DISTANCE

Because the effect of the reflections and externally induced noise is negligible, the expression (2) in Chapter 1, under subhead 1.13, takes the following form:

$$K = SX/0.9S$$

For example, assuming cable attenuation 15 dB/100 m, S = 1 V and cross talk (through connectors) attenuation is 20 dB (10 times), we obtain

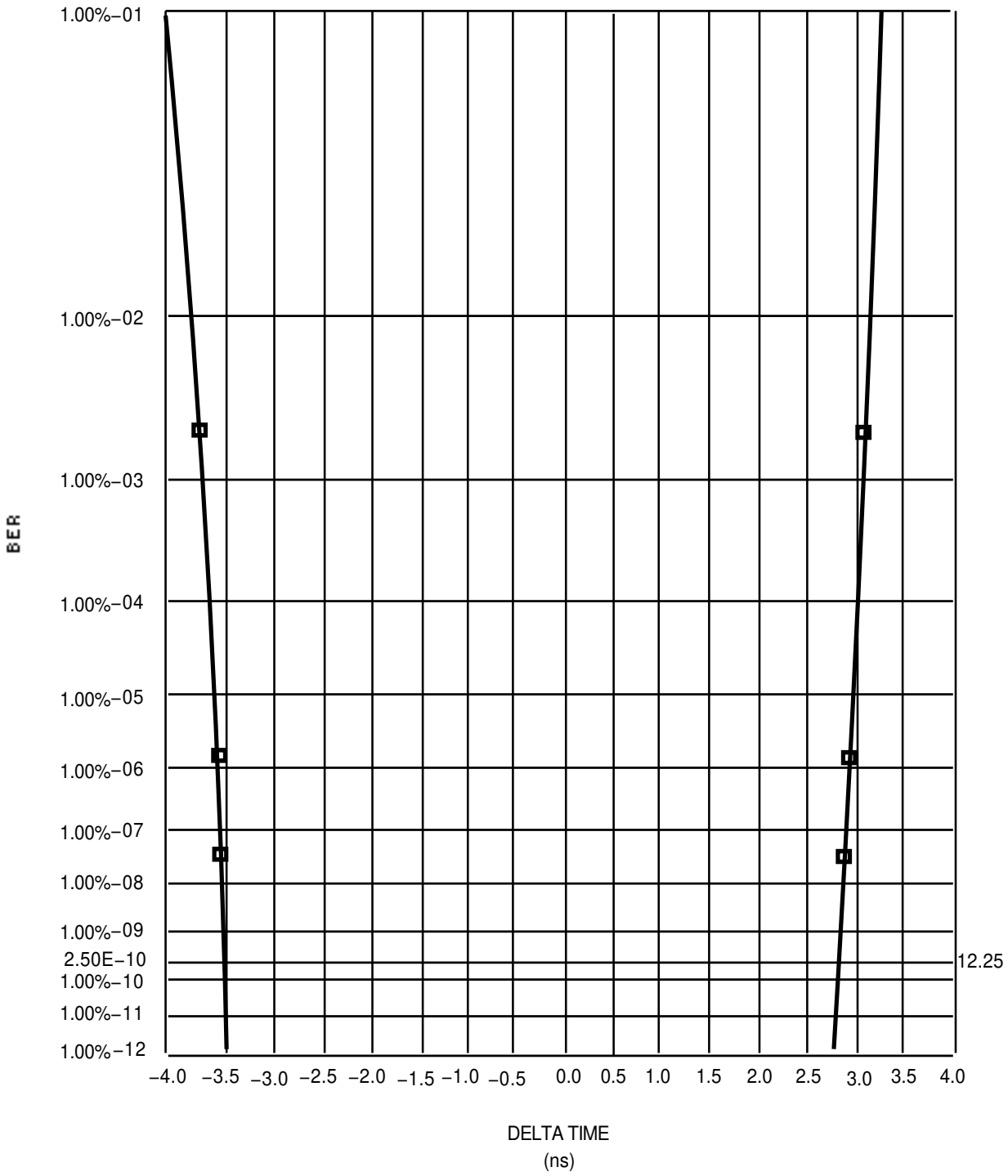
$$K = [1(1/10)]/0.9 * 1 \ = 0.11 \ (19.1dB)$$

The achievable distance is:

$$19.1/15 = 127 \ m$$

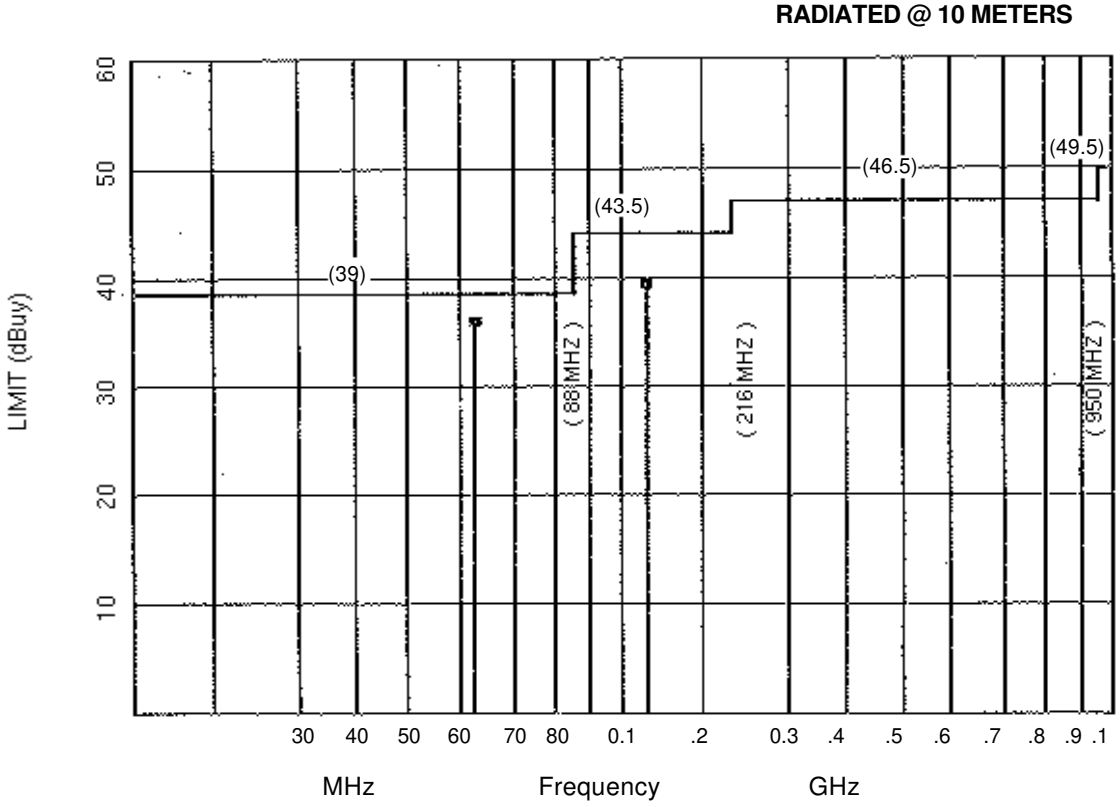
BER (bit error rate) as a function of time deviation from the optimal sampling position (window opening as a function of BER) at a distance of 100 m is shown in Figure 3-5.

Figure 3–5: Window Opening as a Function of BER



Results of emission testing, together with the FCC limit (solid line), are shown in Figure 3-6.

Figure 3–6: Results of Emission Testing, with FCC Limit



CHAPTER 4

UNSHIELDED TWISTED PAIR (UTP) LINK

4.1 MULTILEVEL MODULATION

When transmitting FDDI over DIW, the transmitted spectral width needs to be decreased. The smaller the transmitted spectrum, the lower the signal attenuation and subsequently the distance is larger; the smaller the bandwidth, the lower the cross talk, and again the distance is larger.

Furthermore, the smaller the spectrum, the lower the electromagnetic emission. The lower the emission, the larger the permissible driving voltage and the larger allowable attenuation in-line and, consequently, the larger the achievable distance. Therefore, the major idea is to compress the spectrum.

FCC regulations determine the breaking point in spectrum width requirements. Refer to Figure 1-6. This point is 30 MHz; below this point the FCC-imposed limitations are very much relaxed.

10 base-T as well as 16 Mb/s IBM token ring are employing spectrum width that is about 30 MHz.

To decrease the high-frequency spectral content, it is worthwhile to remain in the baseband area. The use of a carrier will actually move signal spectrum to a higher frequency range (although the spectrum width might be smaller).

The spectrum decrease—while remaining in baseband area—can be achieved by increasing the number of amplitude levels. Specifically, to decrease spectrum by n times with the multilevel modulation, one has to employ 2^n levels.

If what is called Modified Duobinary Modulation (MDB) is employed, the required number of levels decreases by 1, and is now: 2^{n-1} levels.

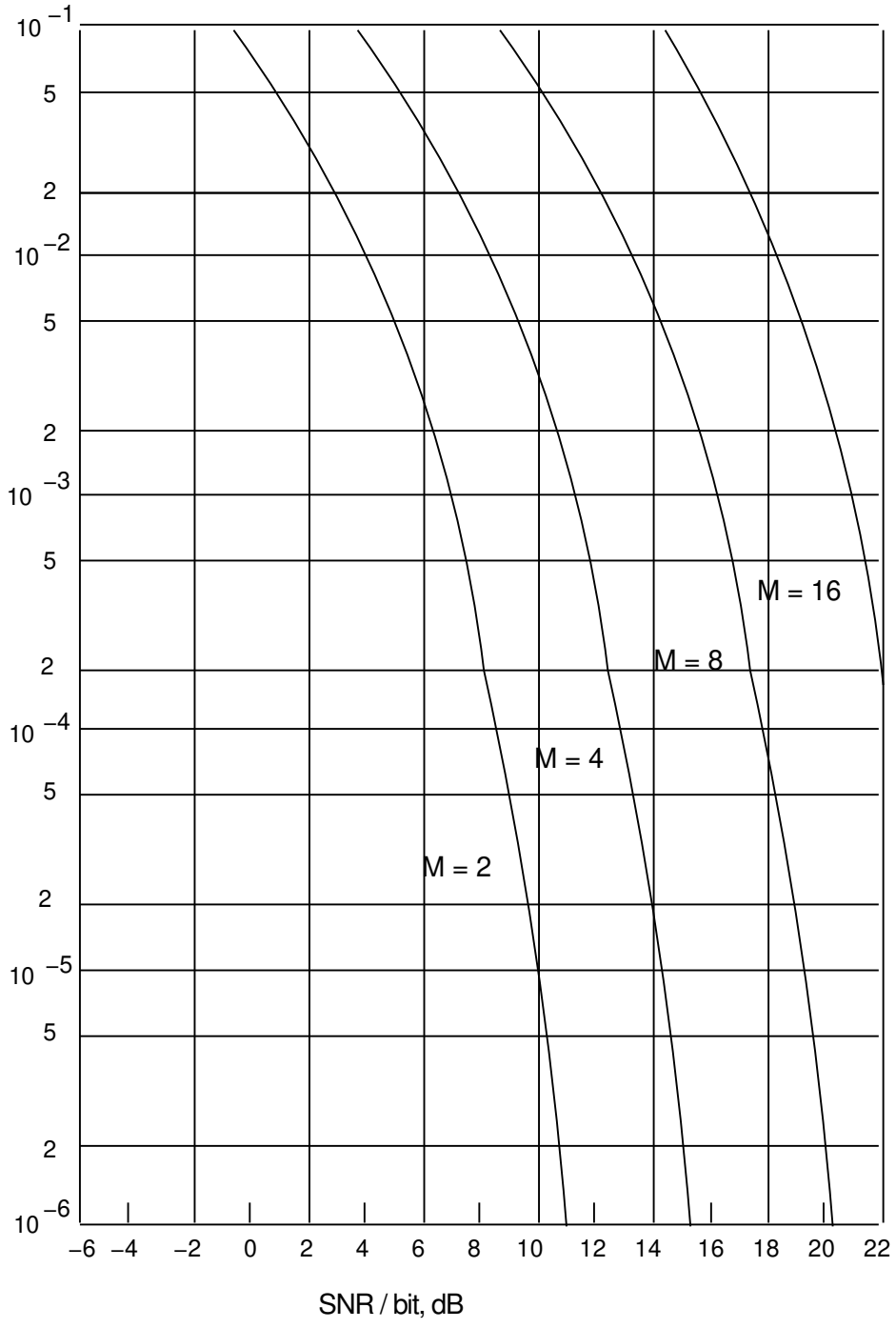
For example, to halve the spectrum (n = 2), 4 - 1 = 3 levels are needed.

To decrease spectrum by 3, 8 - 1 = 7 levels are needed; to compress spectrum by 4, 15 levels are needed, and so forth.

However, when the number of levels increase, noise immunity, which is determined by a distance in voltage between the levels, is lost.

In terms of average signal-to-noise ratio per bit, based on Gaussian type noise, the loss is approximately 4 dB for M = 4 (number of levels), and for large values of M, the penalty approaches 6 dB for every factor of 2 increase in M. Refer to Figure 4-1 [15].

Figure 4–1: Probability of an Error for Pulse Amplitude Modulation (PAM)



4.2 COMPARISON OF DIFFERENT APPROACHES FOR TRANSMITTING FDDI OVER UTP

Figures 4-2 [16], 4-3, and 4-4 show FDDI UTP spectrums for different approaches. Note that the original FDDI spectral width is in the order of 100 MHz, as shown in Figure 4-5.

The difference between methods is in modulation format that results in different degrees of achieved spectrum compression. The time diagrams are shown in Figure 4-6.

Figure 4–2: PSD of Copper Distributed Data Interface Line Codes Without Filtering

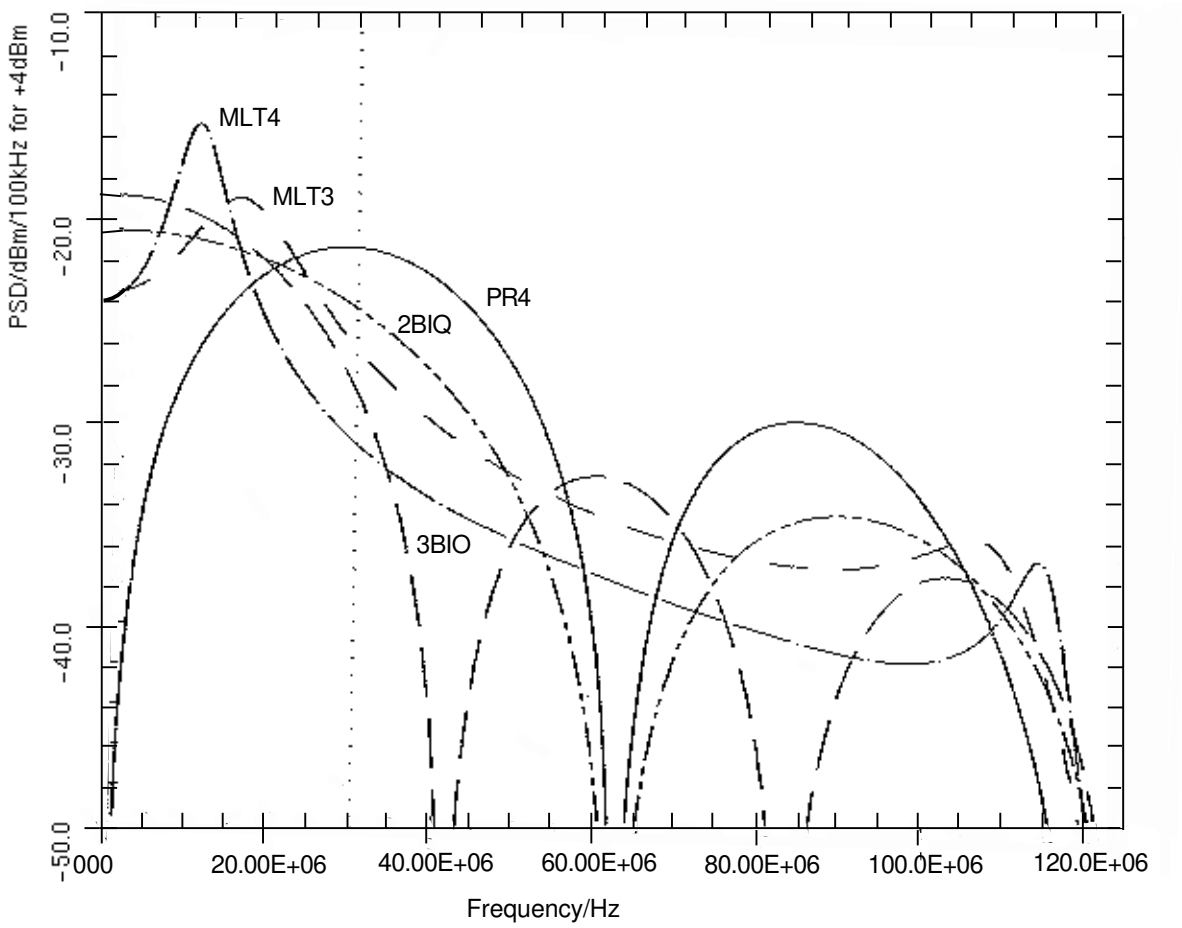


Figure 4–3: FDDI UTP Spectrum

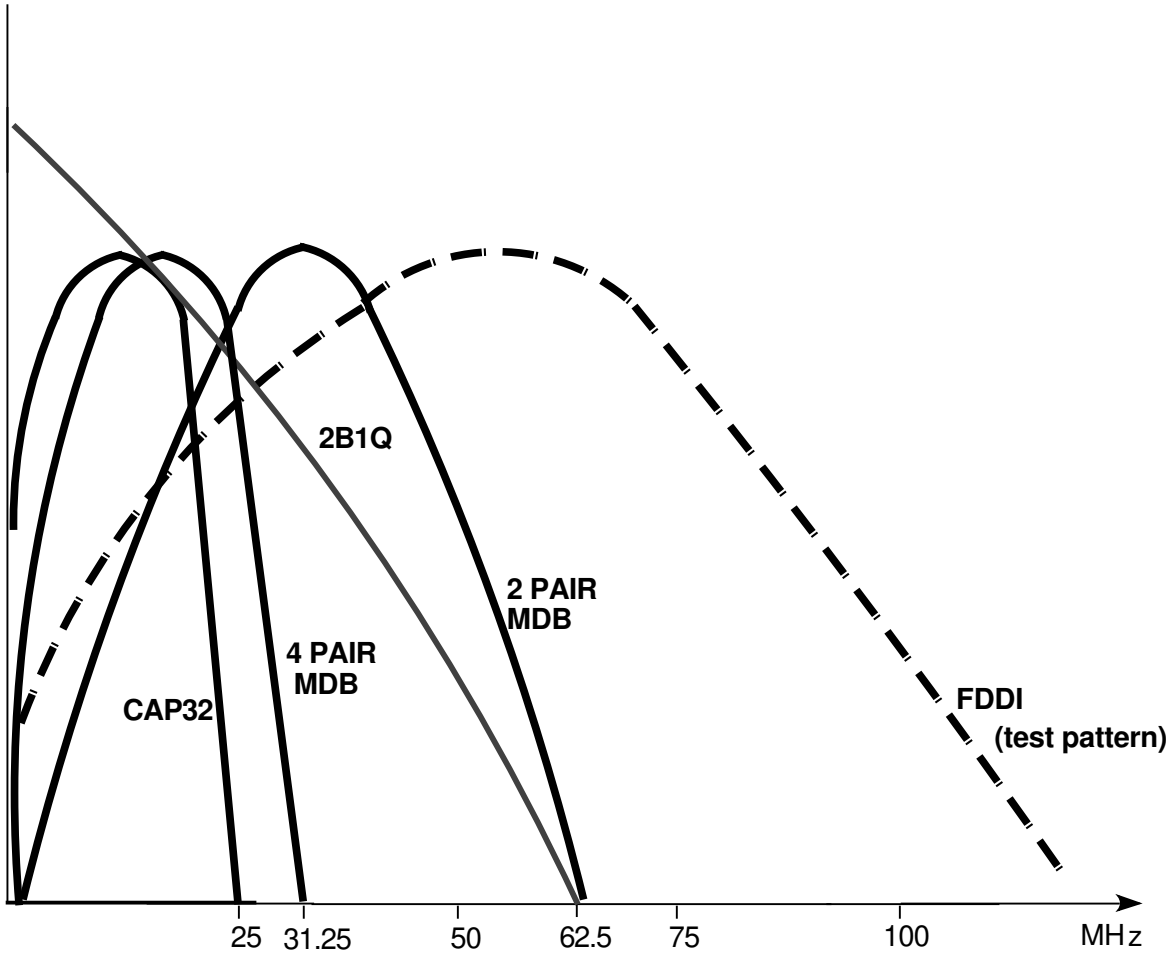


Figure 4–4: Original FDDI Spectral Width

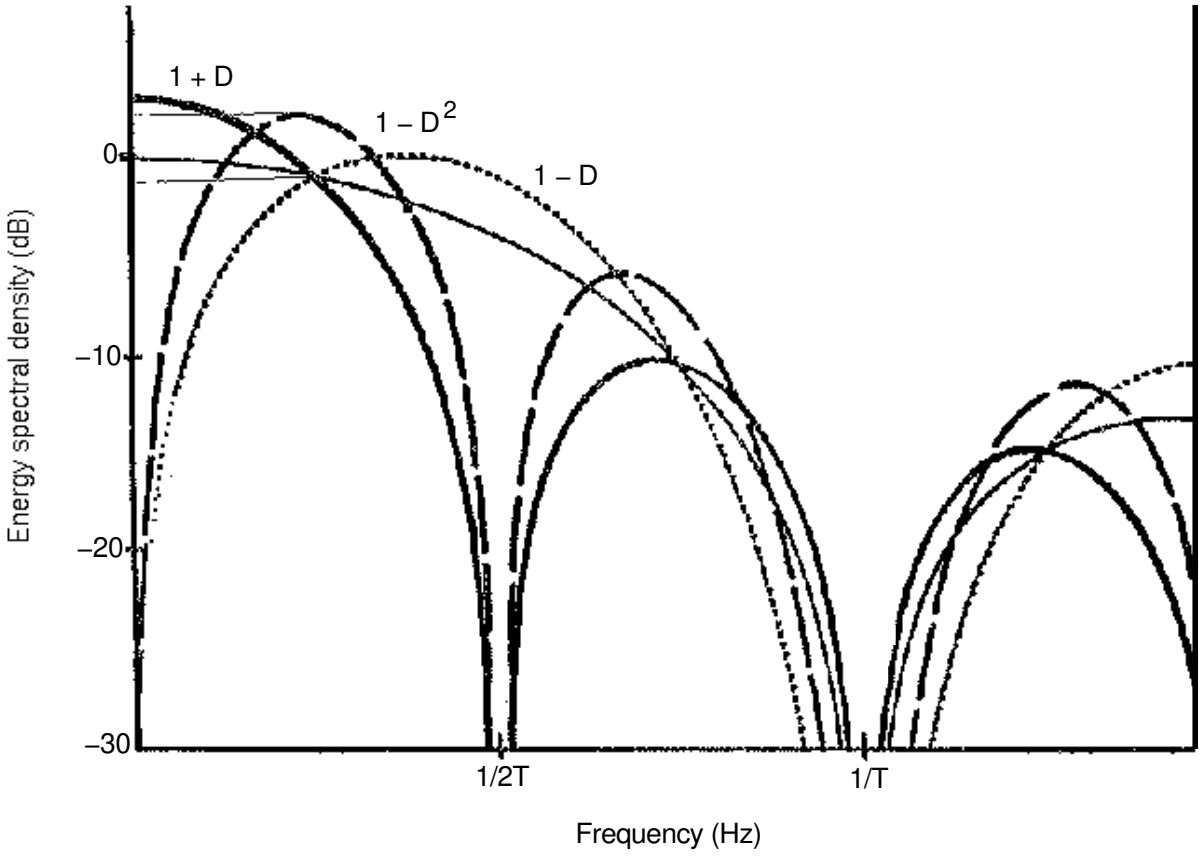


Figure 4–5: FDDI UTP Spectrum

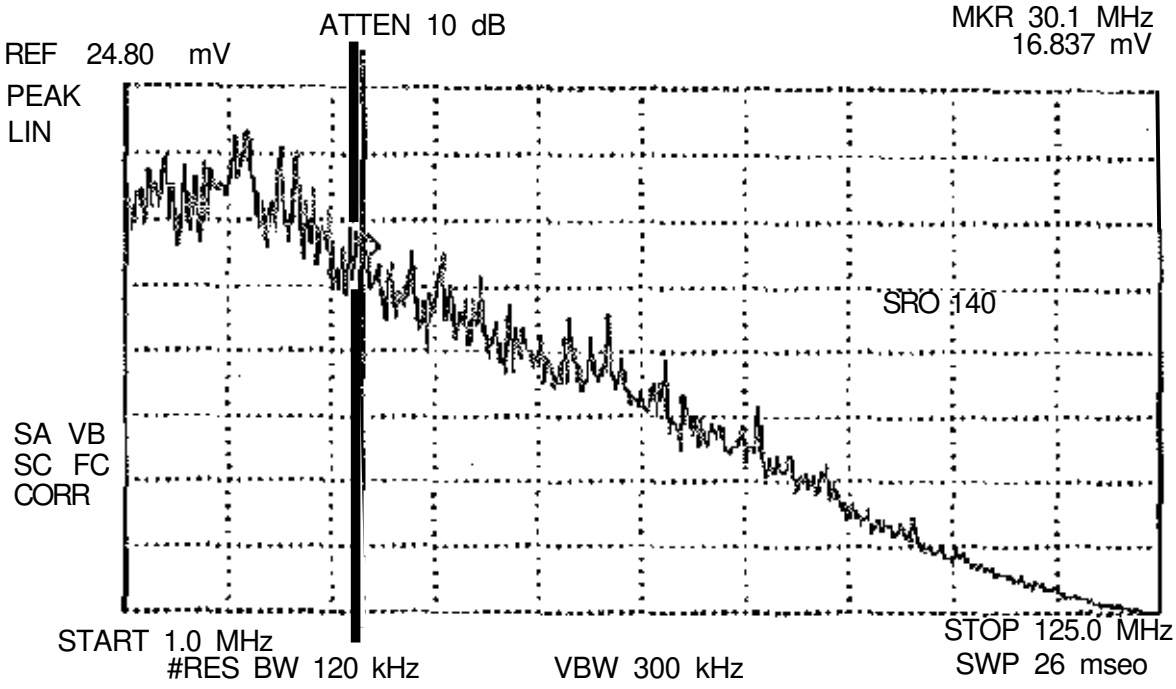
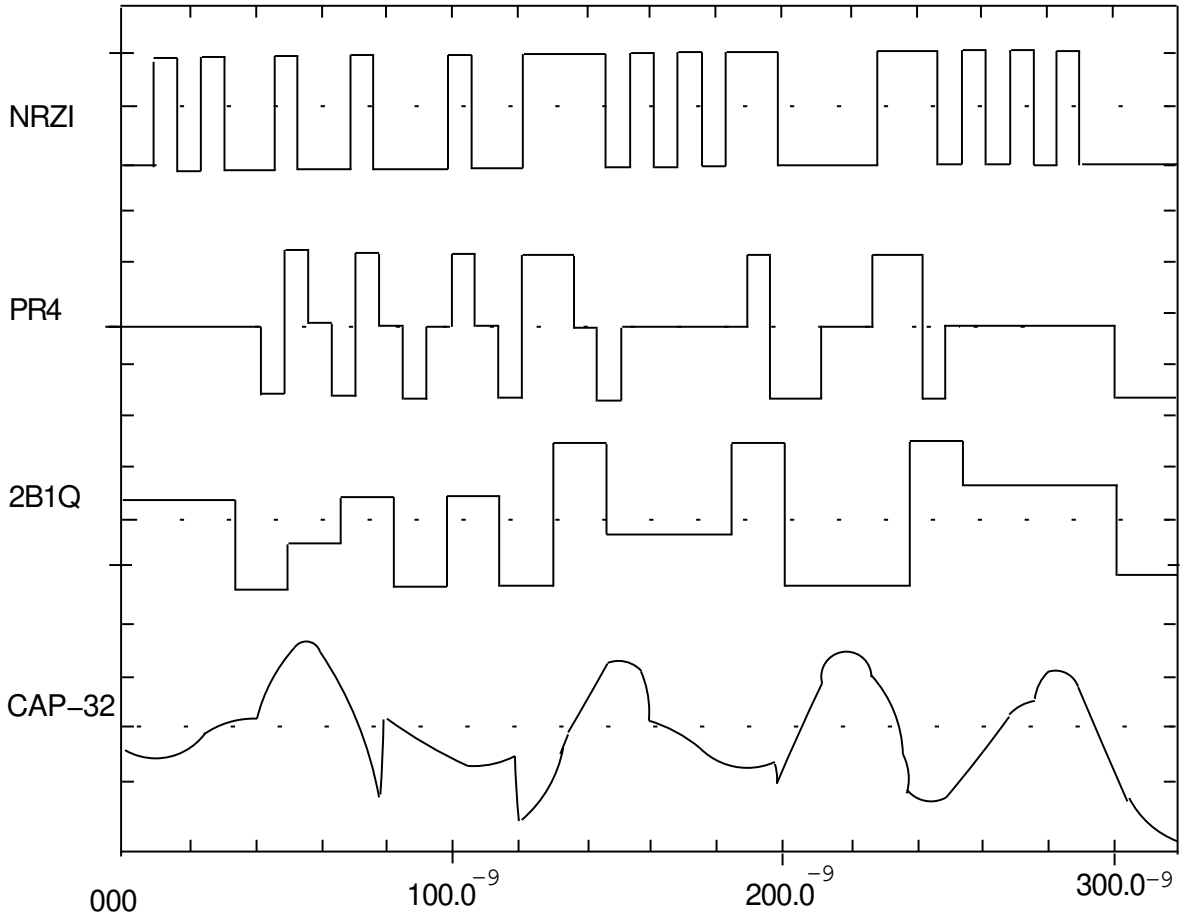


Figure 4–6: Prospective Line Codes in the Time Domain



4.2.1 CAP-32

Using this method, signals are transmitted over two channels that are separated by providing 90-degree shift between proper signals. This shift is actually carrierless phase division. Theoretically, signals in both channels can be separated in the receiver without interference to each other. Refer to Figure 4-7 [17].

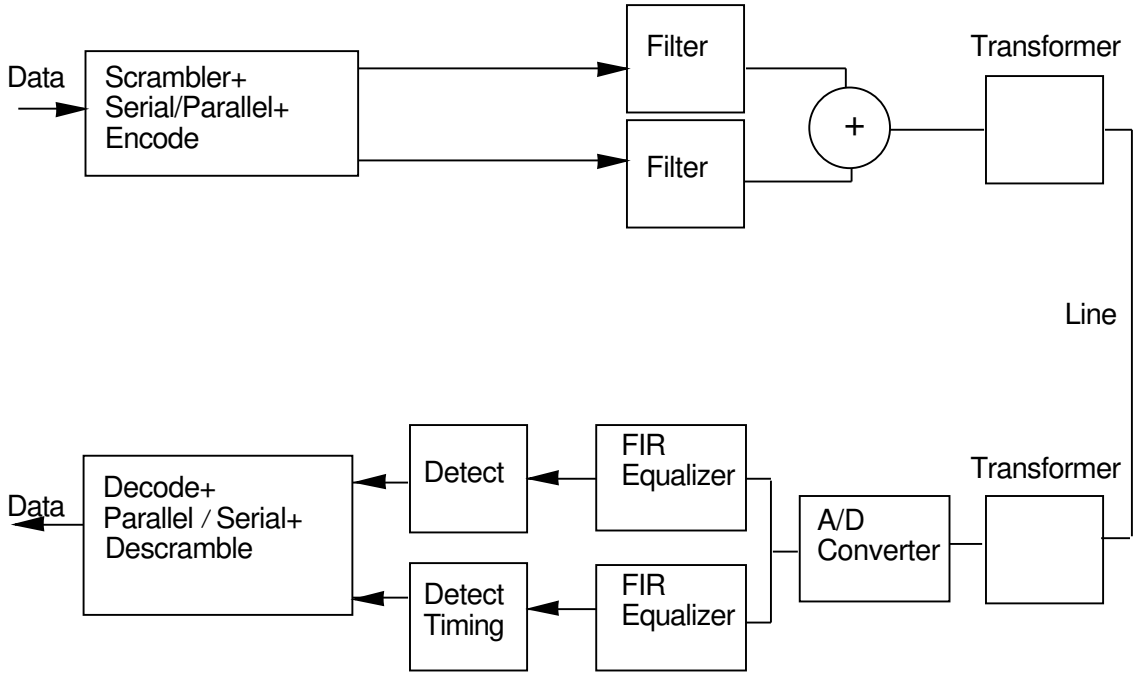
Implementation of carrierless phase division implies the development of an all-pass filter with 90-degree phase shift for all frequencies (Hilbert filter) in a large frequency range, from 1 to 25 MHz, which gives 25/1 = 25.

To decrease signaling speed to 25 Mb/s or 5 times, one has to transmit signals with five or six amplitude levels. Using something similar to partial response signaling, which involves bit overlapping, transmission can be achieved with 5-level signals. The noise immunity (distance in voltage between the levels) for each of the 6-level signals is less by 8 dB, and for 5-level signals, by 6 dB, if compared with a channel with 3-level PR-4 (MDB) signals.

There are two sources of the cross talk in this scheme: imperfection of Hilbert filters and cable itself. This fact, together with 5- or 6-level modulation format, imposes strong restrictions on the cross talk level, which cannot be met without a cross talk suppressor.

This format, after filtering, has all the power located below 30 MHz and therefore is very effective from the FCC standpoint.

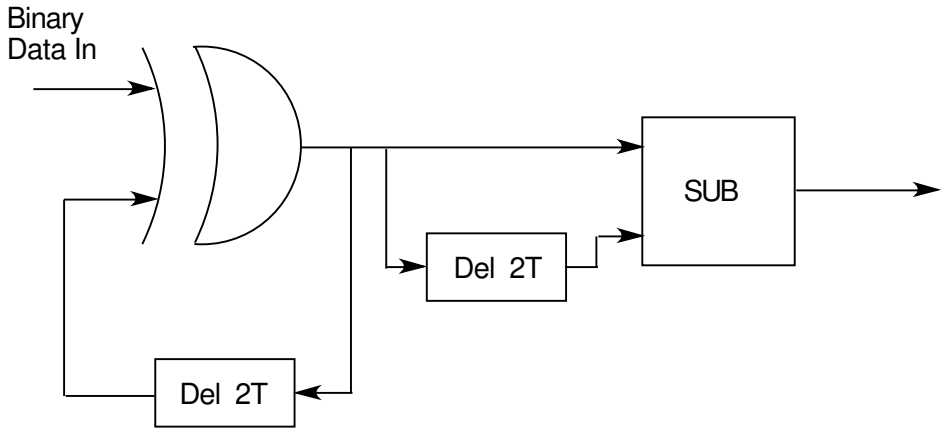
Figure 4–7: CAP-32 Transceiver Structure



4.2.2 PR-4

PR-4, or modified duobinary (MDB) code, is a 3-level format that is allowed to shrink the spectrum about 2 times. Using two pairs for transmission in each direction, the effective spectrum is below 30 MHz [18]. The PR-4 encoder shown in Figure 4-8 includes a delay circuit, which interposes a delay equivalent to two data periods, and a subtractor circuit, in which the delayed binary data signals are subtracted from the undelayed incoming binary data symbols to produce a 3-level duobinary code signal.

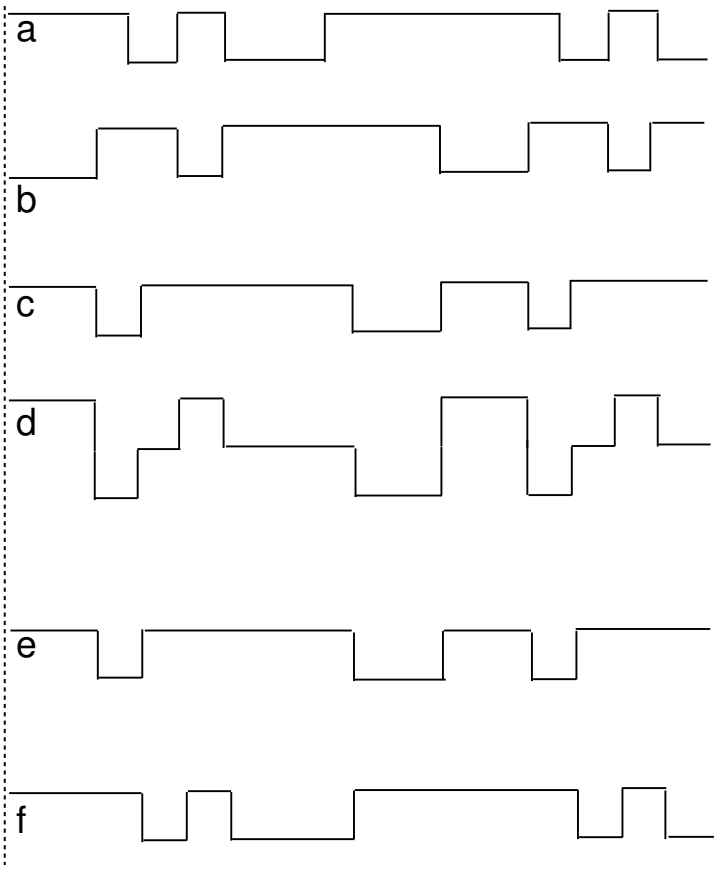
Figure 4–8: MDB Encoder Circuit



To make the output level at the detector directly indicate the original data without comparison to the previous sample value, a precoder is employed. The precoder also serves as a Gray encoder. The precoder includes 2-bit delay and an exclusive-OR gate. One input to the exclusive-OR gate is the incoming binary data. The delay circuit takes its input signals from the output of the exclusive-OR gate and provides the second input to the exclusive-OR gate. The output of the exclusive-OR gate provides precoded binary signals to the subtractor. Two delays are shown in Figure 4-5 for explanation purposes: one delay circuit is sufficient and can be used in both precoder and encoder.

Figure 4-9, a through f, provides an example of the MD encoding and decoding process. Figure 4-9a is an illustrative waveform of a binary signal at the input of the modulator. Figure 4-9b is the waveform of a corresponding signal output from the delay circuit and applied to the other input of the exclusive-OR. Figure 4-9c is the resultant output of the exclusive-OR gate; it represents the exclusive-OR of the signals shown in Figures 4-9a and 4-9b.

Figure 4–9: Modified Duobinary (MDB) Format



It should be noted that the waveform shown in Figure 4-9b is delayed by two time cycles with respect to the output signal of Figure 4-9c, which represents the precoded binary input signal.

The subtractor circuit subtracts the delayed input signal, which is the same as the one shown in Figure 4-9b, from the undelayed input signal shown in Figure 4-9c. The result is the 3-level signal of Figure 4-9d, which is then filtered in the low-pass filter and transmitted onto one of the twisted pairs in a form similar to the waveform shown by the broken line in Figure 4-9d.

In the receiver, levels of +1 and -1 in the transmitted data are translated into a logical 1 and a 0 level is translated into a logical 0, as shown below:

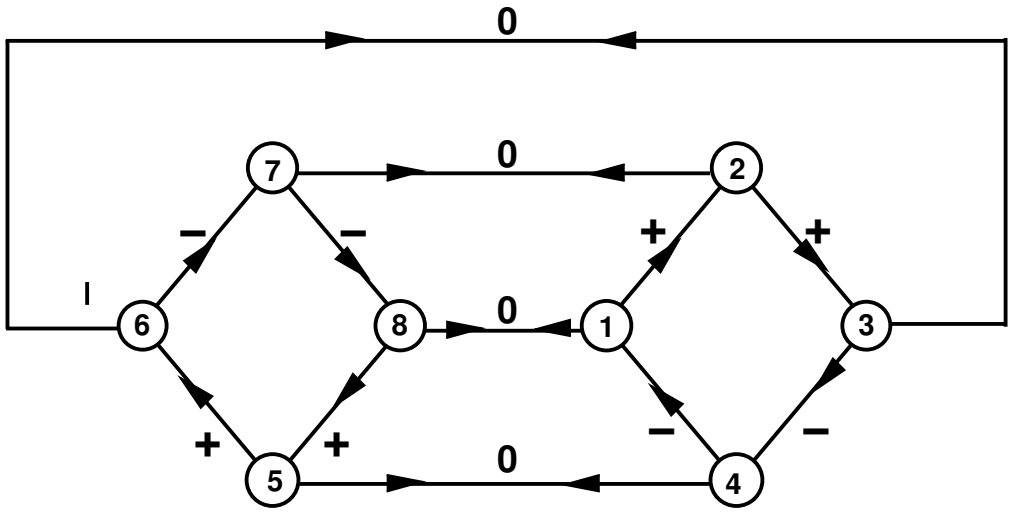
```
TD:  +1  +1  −1   0  +1   0   0  −1  −1  +1  +1  −1   0  +1   0
DD:   1   1   1   0   1   0   0   1   1   1   1   1   0   1   0
```

The resulting decoded data is shown in Figure 4-9f, which is the same as the waveform of Figure 4-9a input to the precoding section of the modulator.

Therefore, at the receiving end, the decoder is simply a rectifier, which translates all nonzero levels into logical 1.

In general, the MDB signals follow the state diagram shown in Figure 4-10 [19].

Figure 4–10: State Diagram for MDB

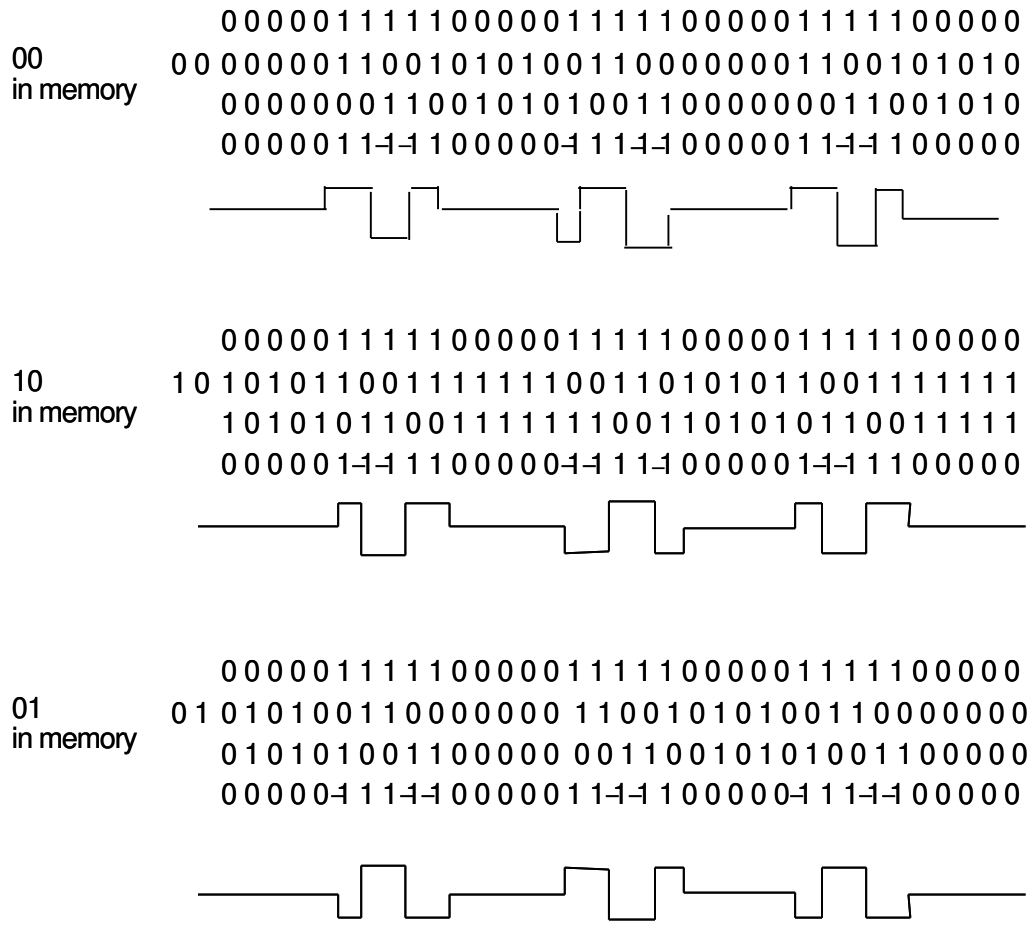


Input / Output Notations:

0 / 0	0
1 / +1	+
1 / –1	–

The spectrum of PR-4 is shown in Figure 4-4 ($1\text{-}D^2$). Remaining from the past history, two bits effect the starting phase (or starting point on the state diagram). If 00 is in memory, we start from point 1 of the state diagram, if 10 is in memory, we start from point 2 and if 01 is in memory, we start from point 4. Figure 4-11 illustrates memory effect for encoding Halt Line State.

Figure 4–11: Memory Effect



The FDDI line states MDB encoded are shown in Figures 4-12 through 4-15.

Figure 4–12: FDDI Line States—Halt (H)

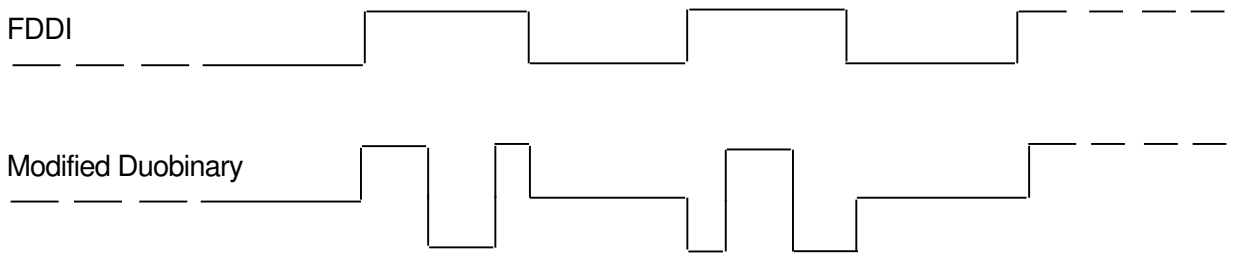


Figure 4–13: FDDI Line States—(MLS)

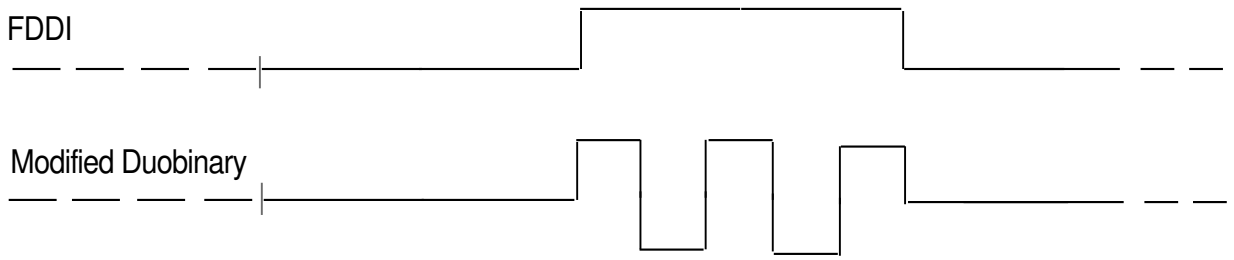


Figure 4–14: FDDI Line States—Idle

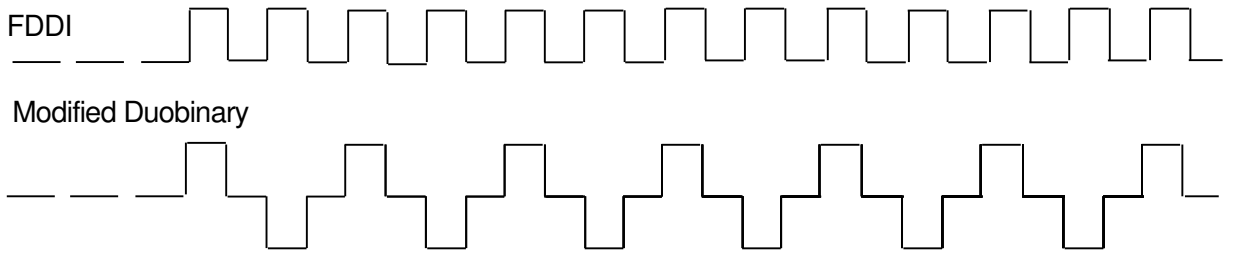
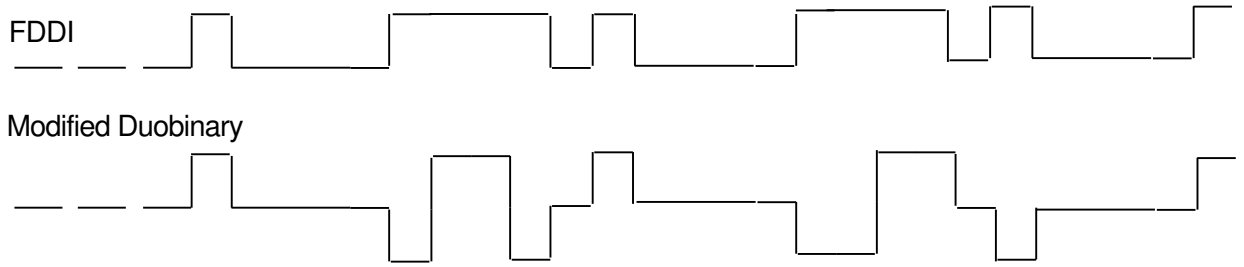
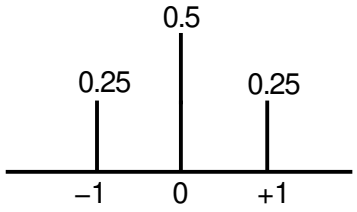


Figure 4–15: FDDI Line States—(J,K)



In general, an error in 3-level modified duobinary signal (MDS) occurs as misinterpretation between adjacent levels. Then, due to the Gray encoder, an error results in only one bit error in the transmitted binary sequence. Note that both extreme levels are decoded into the same binary 1, and therefore misinterpretation between extreme levels does not cause an error in the demodulated pulse train.

Actually MDS has a triangular distribution, as shown below.



Because the MDB signal is balanced, the number of +1s equals the number of -1s.

If, instead of two of 2T delays in the encoder (Figure 4-8), just T delay is used; the duobinary format (DB) is obtained. Figure 4-4 compares the spectrums MDB ($1 - D^2$) and DB ($1 + D$). At 30 MHz, DB has smaller spectral density by about 3.5 dB

4.2.3 2B1Q

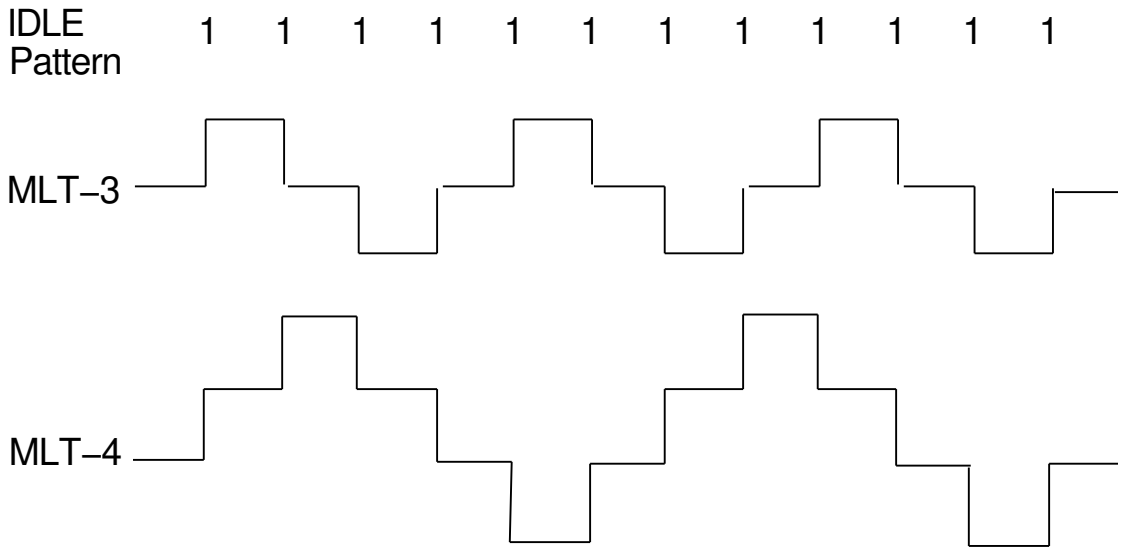
2B1Q (2 binary, 1 quaternary) is 4-level pulse amplitude modulation (4-PAM), accepted by ISDN for North America. Refer to Figure 4-6. The code converts blocks of two consecutive binary bits into a single 4-level pulse for transmission. The information rate is double the baud rate.

Contrary to MDB, the bandwidth reduction in the 2B1Q is achieved without using partial-response signaling. Therefore, the number of levels 4 versus 3 for PR-4. More levels result in lower noise immunity by about 3.5 dB.

4.2.4 MLT-3 and MLT-4

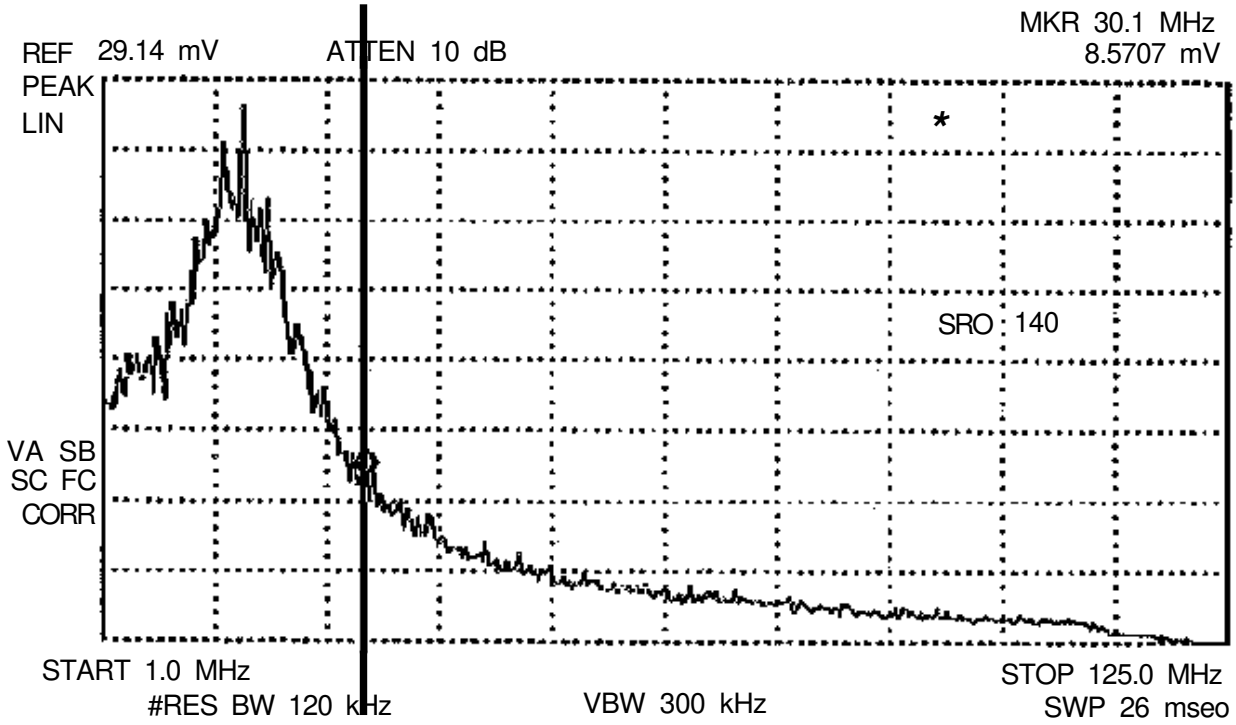
MLT-3 is a 3-level format [20] and MLT-4 is a 4-level format. Each 0 input bit causes the current output to be repeated, while each 1 input bit causes the output sequence to be advanced by one level. Refer to Figure 4-16. The spectrum of MLT-3 is shown in Figure 4-17.

Figure 4–16: MLT-3 and MLT-4 Formats



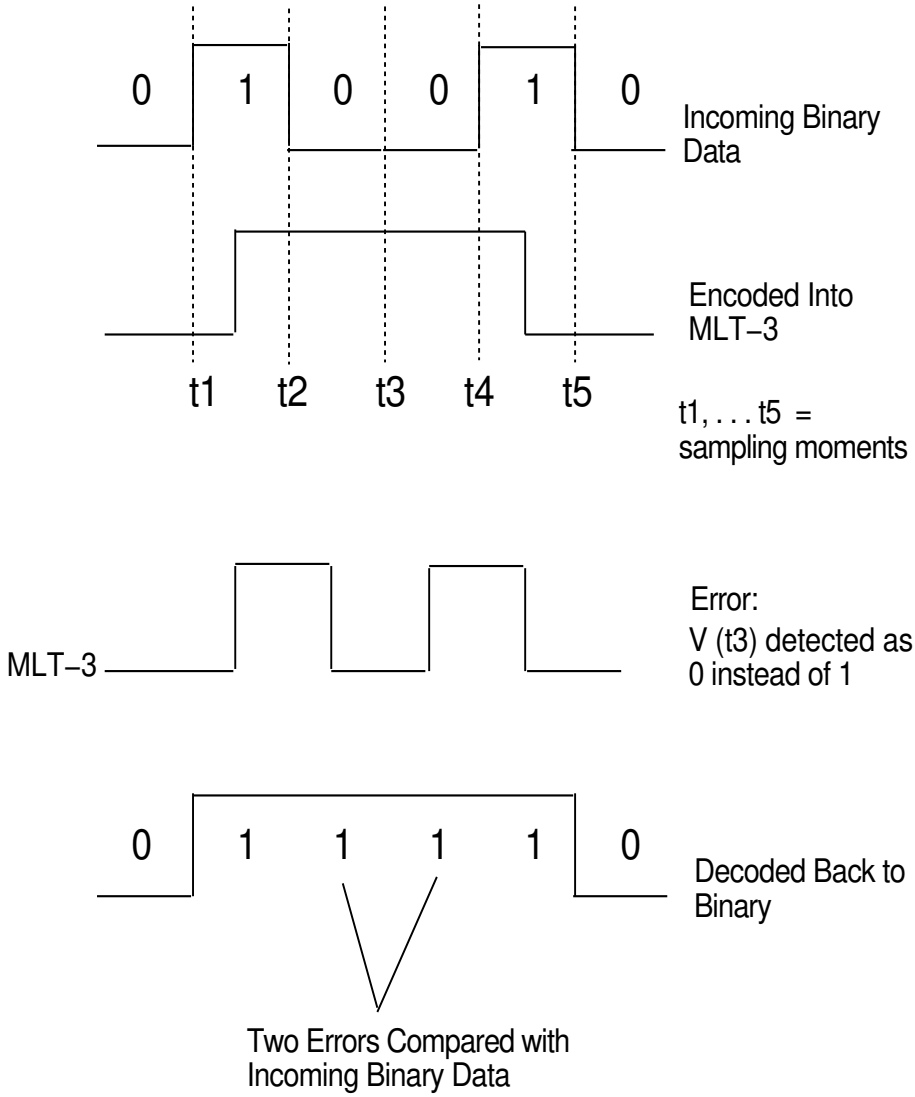
Refer to Figure 4-15. Both formats have strong dc and low frequency components; the spectrum occupies all the 125-MHz range, although just about 16% (for MLT-3) and 7% (for MLT-4) of all the signal energy is beyond 30 MHz.

Figure 4–17: MLT Spectrum



The major drawback is that both of these formats cause error propagation, as shown in Figure 4-18, and require dc (or almost dc) coupled receivers. The run length can be unacceptably long. This drawback makes equalization more difficult than formats with 0 spectral density around dc. It will also result in strong baseline wander.

Figure 4–18: Error Multiplication in MLT-3



4.2.5 3B10

3B10 is actually 8-level PAM. The spectrum main lobe has 0 at 40 MHz and only 8.6% of the total power is beyond 30 MHz. However, an 8-level signal looses about 11 dB to the 3-level signals, with regard to distance between levels (noise immunity) and therefore requires cross talk suppression.

4.2.6 QAM-32

This format is similar to CAP-32 except it uses 25-MHz carrier. The signal spectrum occupies 50 MHz frequency range, versus 25 MHz of CAP-32. But the modulation and demodulation are simpler; there is no need for Hilbert filtering.

4.2.7 UTP TRANSCEIVER BLOCK DIAGRAM

Silicon Systems proposed the transceiver configuration shown in Figure 4-19, incorporating multilevel signaling, an adaptive cross talk suppressor, fixed transmitter equalizer, an adaptive receiver equalizer in frequency domain (continuous time), and scrambling [21].

4.3 ENHANCING TECHNIQUE

The following is a review of different enhancing techniques that were proposed to provide an UTP FDDI commercial distance.

4.3.1 ADAPTIVE EQUALIZATION

Adaptive equalization [15] is needed for approaches with large bandwidth and due to unknown and varying cable parameters. A fixed analog equalization does not provide sufficient performance in this case. An adaptive filter-equalizer normally employs a transversal filter, which is a delay line tapped usually at bit duration intervals. Each tap is connected through a variable gain device X to a summing amplifier S, as shown in Figure 4-20.

Figure 4–19: Possible UTP Transceiver Block Diagram

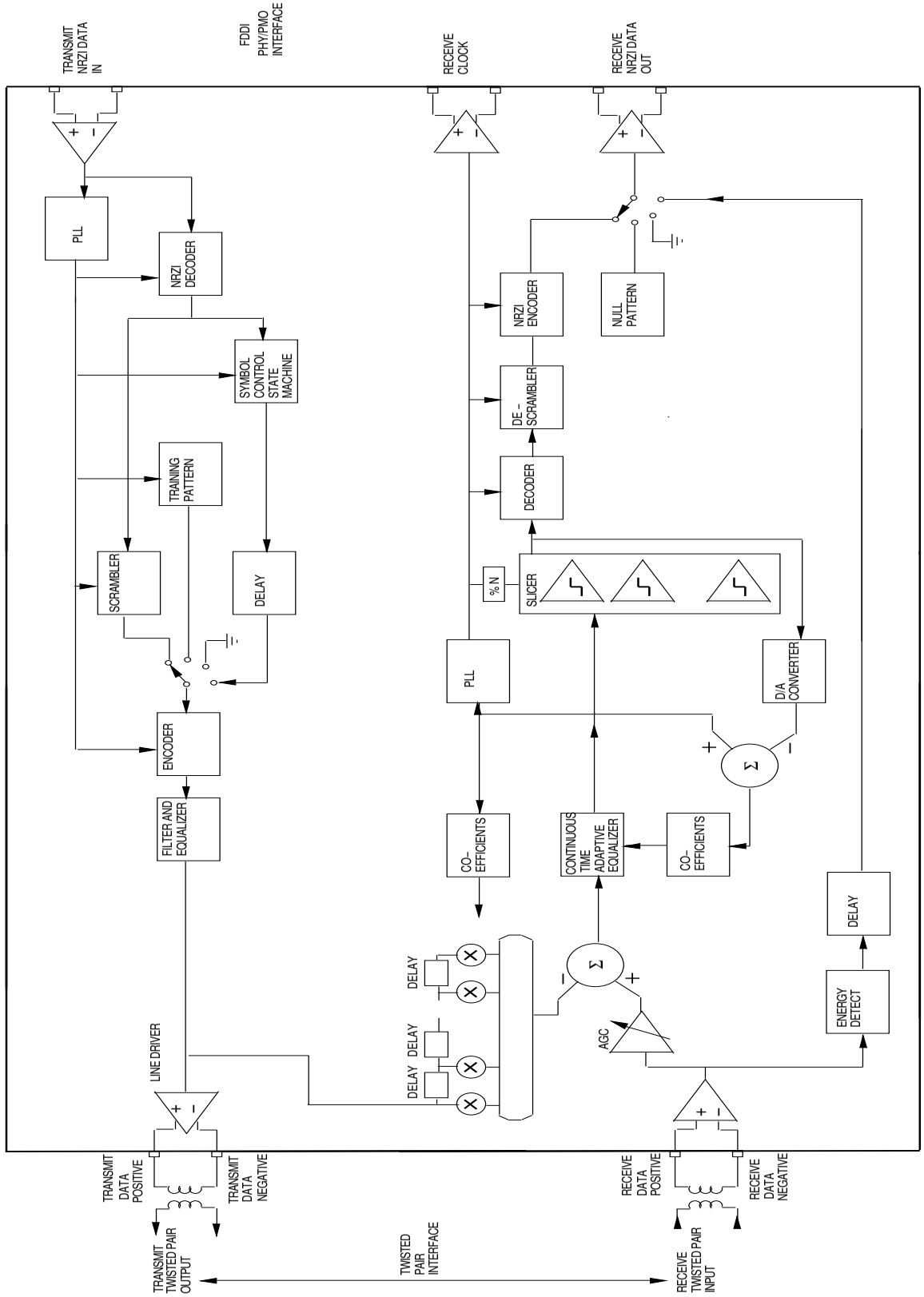
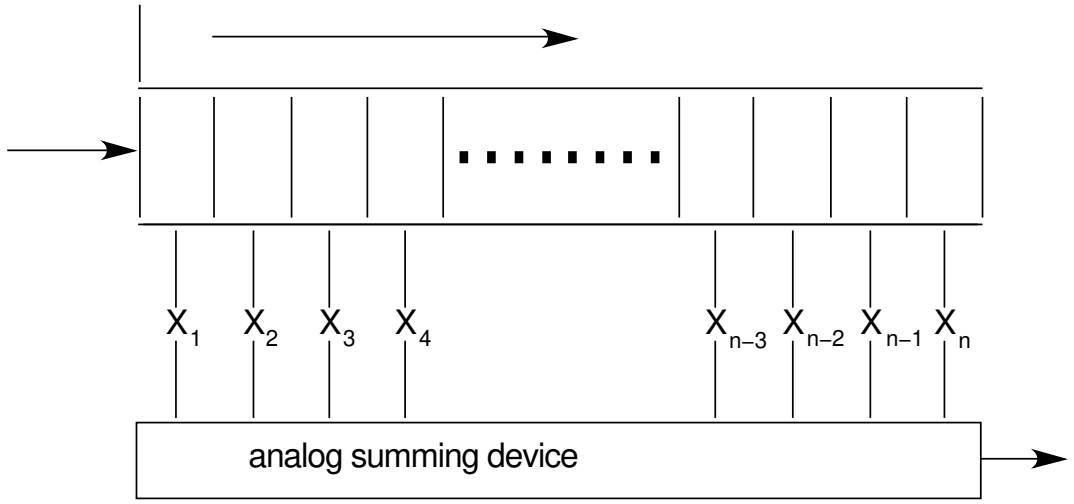


Figure 4–20: Adaptive Filter-Equalizer



By using different tap gains, different transfer characteristics are obtained.

The adaptive filter-equalizer employs iterative techniques to arrive at a specific set of the tap gains.

The iterative technique is based on the fact that many characteristics of the transversal filters are convex functions of the gain coefficients. In other words, they possess global minima and no relative minima. The optimum gain distribution is carried out by choosing arbitrarily a set of tap gain coefficients, then changing them in the direction opposite to its corresponding error component.

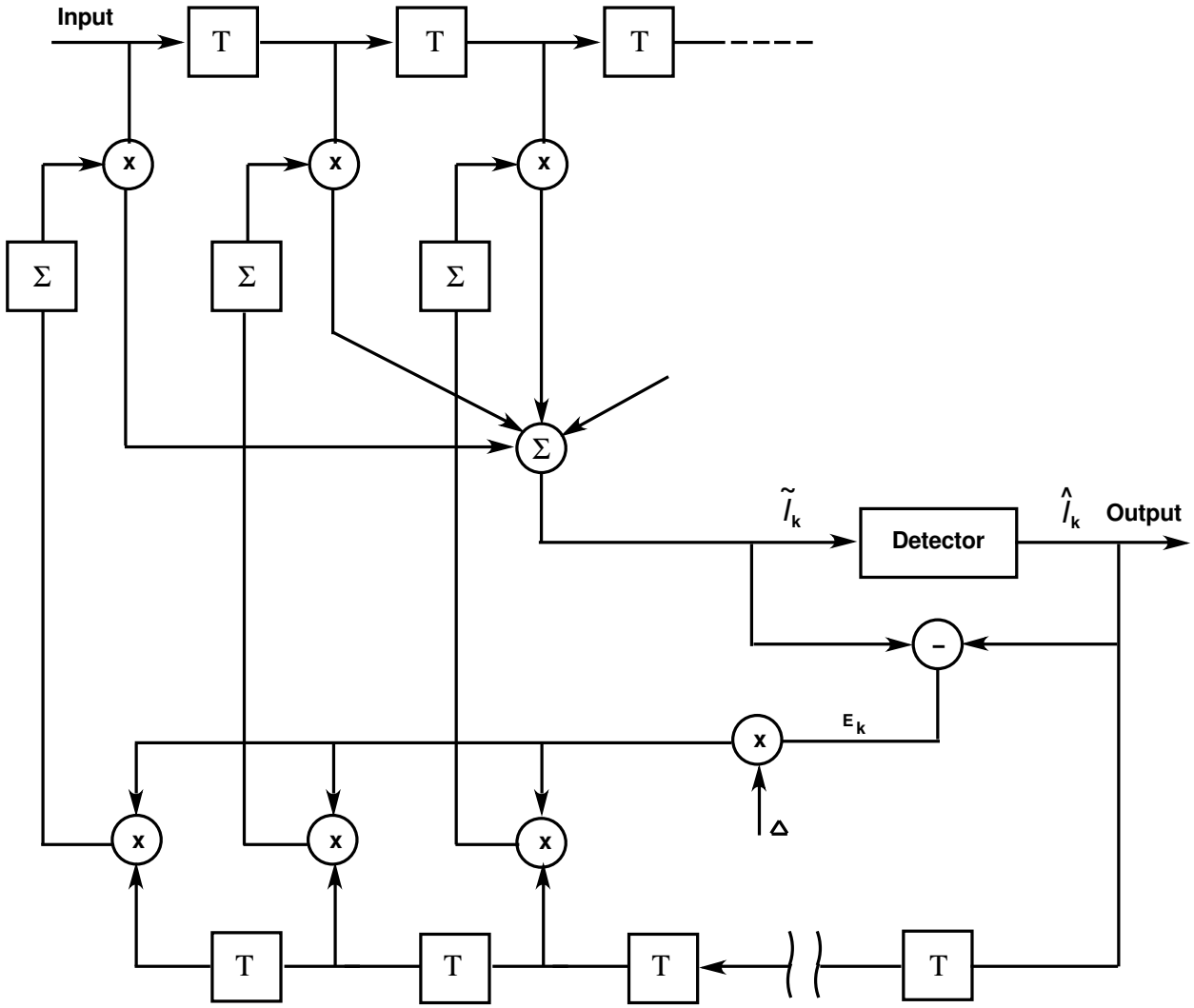
In the traditional application as an adaptive line equalizer, there are three most popular criteria and corresponding algorithms.

4.3.1.1 Zero-Forcing Algorithm (ZFE)

In ZFE, the tap weights are adjusted to force the cross-correlation products of signal (estimation after decision) and the ISI to be 0 or less than an arbitrary small number. Refer to Figure 4-21.

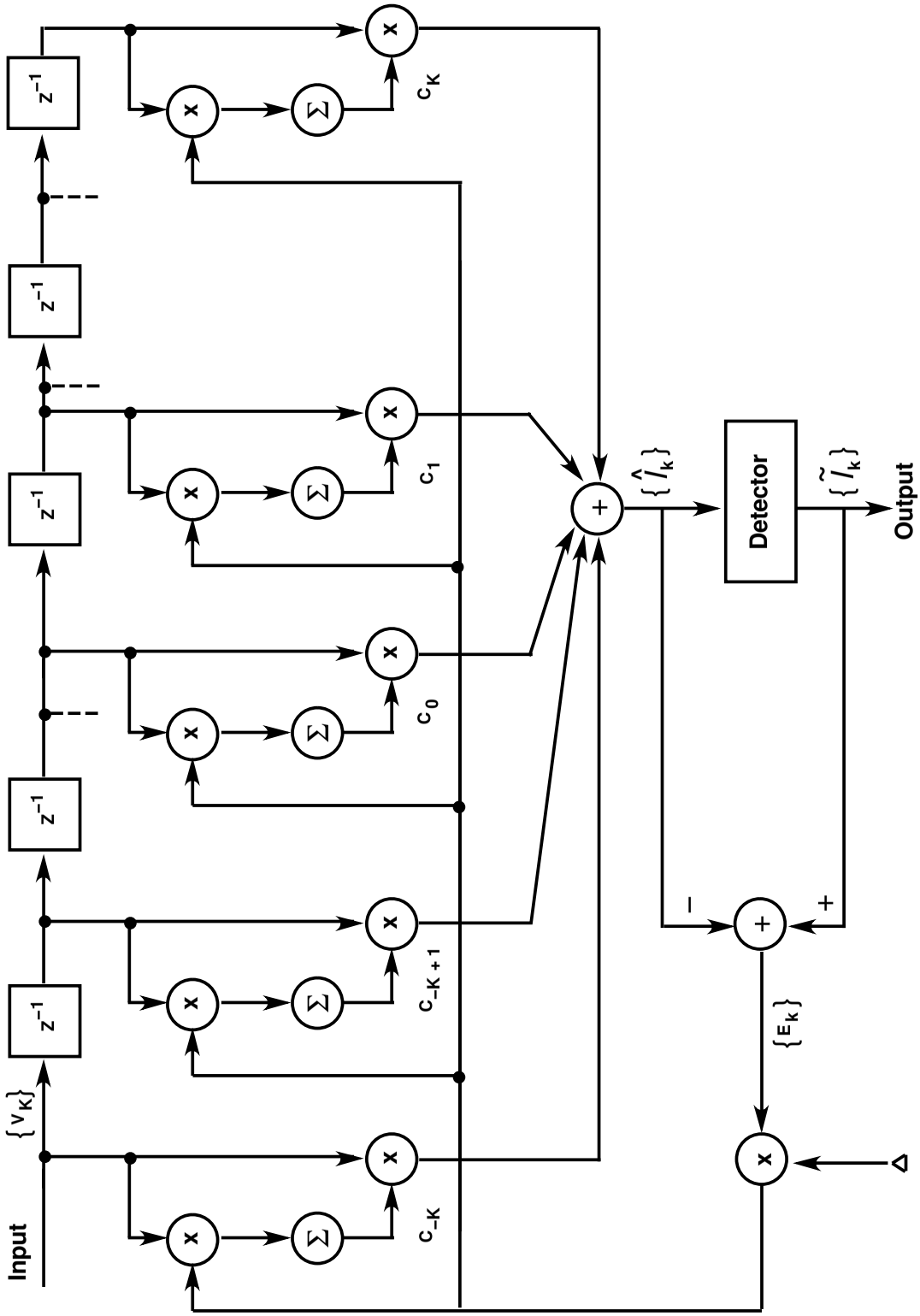
The zero-forcing algorithm starts working only if the eye was not initially completely closed. This condition is often required to send a training pattern for initial adjustments. The second point is that the peak distortion criterion neglects the effect of additive noise. These limitations do not exist if the mean square error criterion (MSE) is used in the adjustment of the equalizer coefficients.

Figure 4–21: Zero-Forcing Algorithm



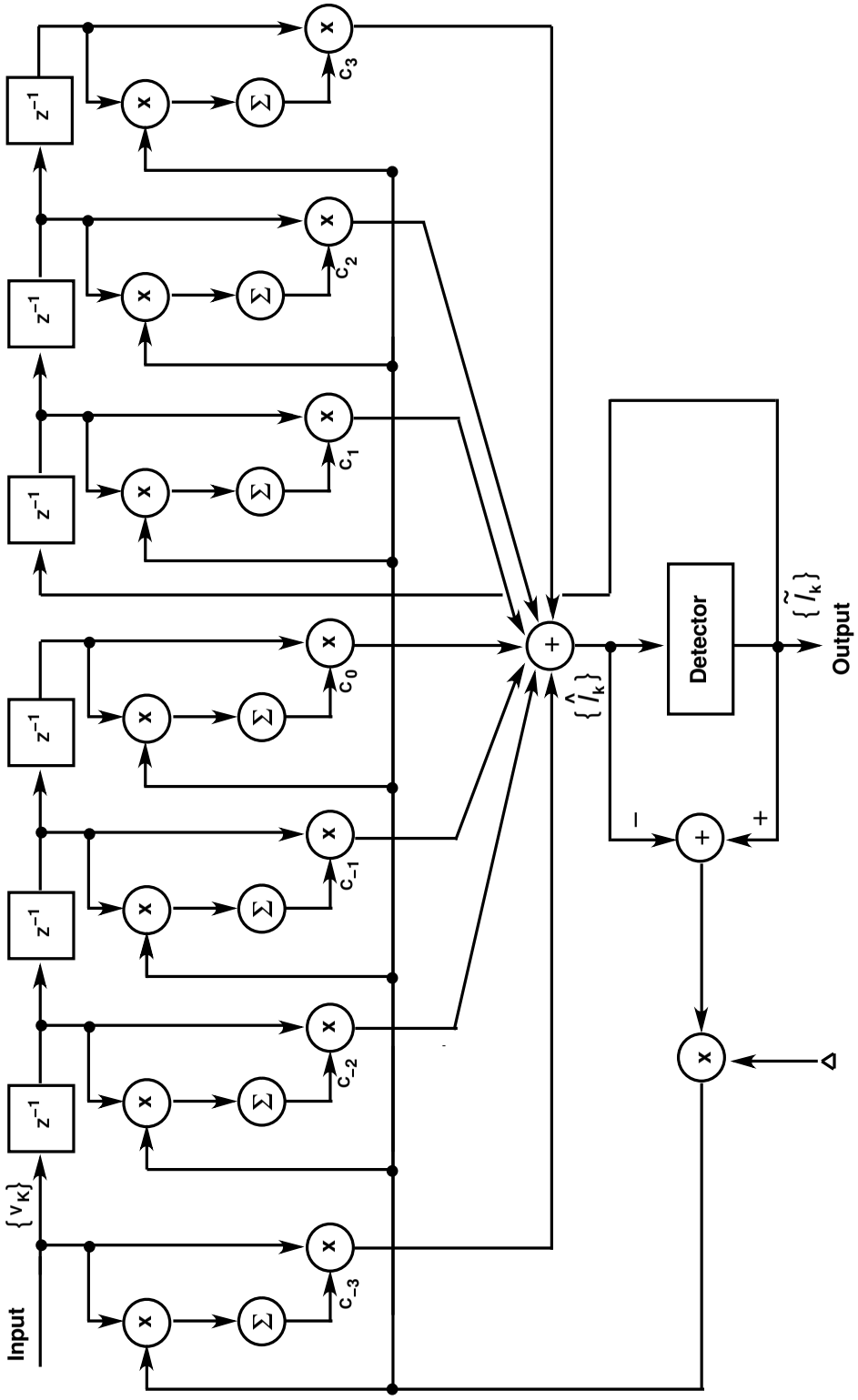
In this method, the ISI and the incoming signal (not the detected pulse train as in the previous algorithm) cross-correlation products are forced to 0. See Figure 4-22. This equalizer will not try to completely equalize the transmission channel if the thermal noise starts to dominate over ISI, giving the optimum weights to ISI versus thermal noise.

Figure 4–22: MSE Equalizer



The decision feedback equalizer provides better performance than ZFE and MSE, but it is at least twice more complex: feedforward (right side) and feedback sections. All adaptive equalizers are normally microprocessor controlled in order to provide right timing— tap gains are adjusted one by one. Only one is adjusted at each time. With 30-MHz bandwidth, each of these equalizers is quite a challenge because many operations must be performed during each bit. See Figure 4-23.

Figure 4–23: Decision Feedback Equalizer

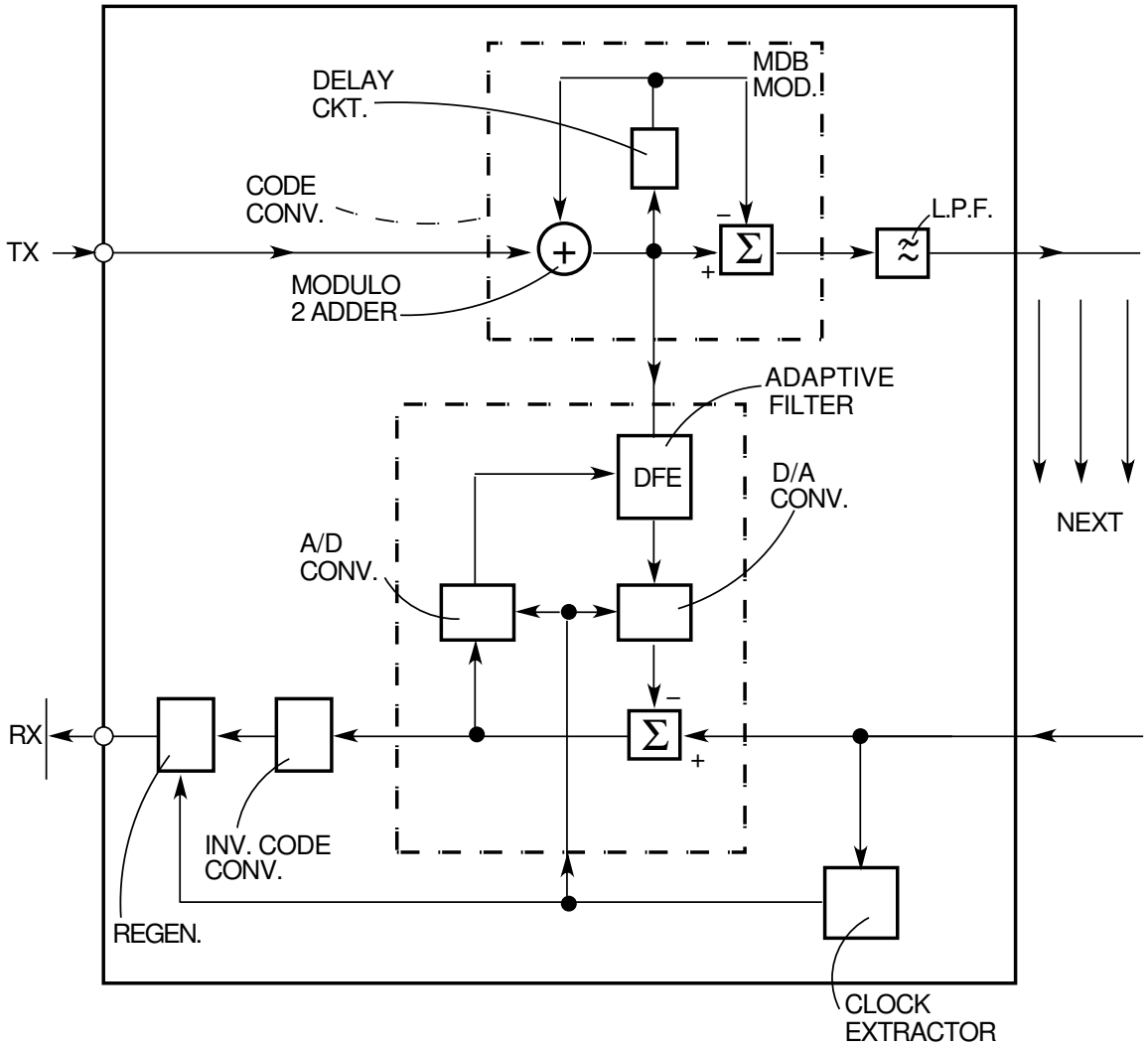


4.3.2 Cross Talk Suppressor

Transmission systems with more than three levels have to employ some kind of cross talk suppressor, which is usually an adaptive filter placed between adjacent transmitter and receiver.

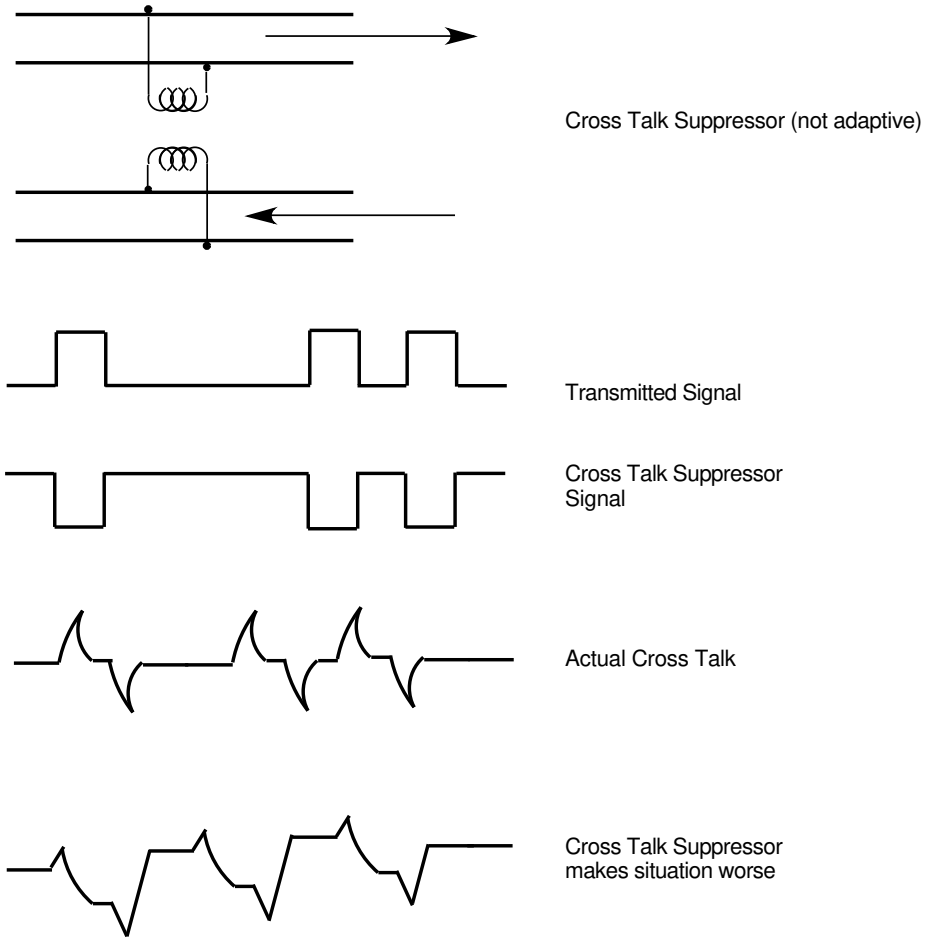
In a cross talk suppressor, a decision feedback equalizer (with proper tap gains) is normally employed. The tap gain coefficients are adjusted in such a way as to create voltage similar to cross talk, but of the opposite polarity, reducing the effect of the cross talk. Refer to Figure 4-24. The cross talk suppressor has to be adaptive due to unknown cross talk transfer functions that depend on the specific cable sample and temperature.

Figure 4–24: Cross Talk Suppressor



If the cross talk suppressor is not adjusted, it can actually make the situation worse. Refer to Figure 4-25.

Figure 4–25: Cross Talk Wave Forms


Cross Talk Suppressor (not adaptive)
Transmitted Signal
Cross Talk Suppressor Signal
Actual Cross Talk
Cross Talk Suppressor makes situation worse

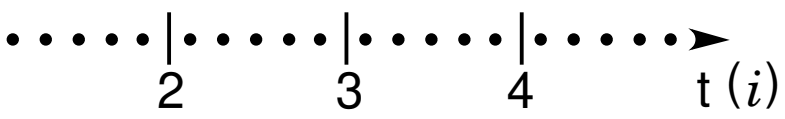
4.3.3 VITERBI ALGORITHM

A receiver employing Viterbi Algorithm (VA) calculates mean square error (MSE). See expression (1) in Figure 4-26. This expression is some derivative of a posteriori probability (AP) for a bit sequence. Minimum in MSE corresponds to maximum in AP. VA provides the decision corresponding to maximum of the AP.

In other words, the receiver selects a sequence which most likely was transmitted based on the waveform, taking into account the previous history.

Specifically, for each bit, the receiver measures the signal amplitude Yi at sampling time i, then it calculates the difference between this value Yi and all the expected amplitude levels Xi—there are m of them. In 3-level MDB, the m = 3 because each symbol is a number selected from the set (0, +1, -1). L is a number of time durations that has effect on sample at i-th interval.

Figure 4–26: Viterbi Algorithm



9 statistics:

MDB levels at point 2

X_1: –1, –1, –1, 0, 0, 0, +1, +1, +1

MDB levels at point 4

X_2: –1, 0, +1, –1, 0, +1, –1, 0, +1

$$\sum_{i=1}^{L} \{ Y_i - SX_i{}^m \}^2 \qquad (1)$$

$$L = 2$$

In other words, the MDB modulator uses 3-bit history (see Figure 4-8). For instance, at step 4 the modulator remembers just current 4 plus 3 and 2. Number 3 has to be remembered because it will be used in the next step, number 5.

The receiver calculates expression (1) for 2 points: 2 and 4, each can have 3 values: -1, 0 and +1. Altogether, it gives 9 statistics.

The receiver will calculate 9, but will remember just the smallest one or two (if there is no clear winner) in regards to each time point.

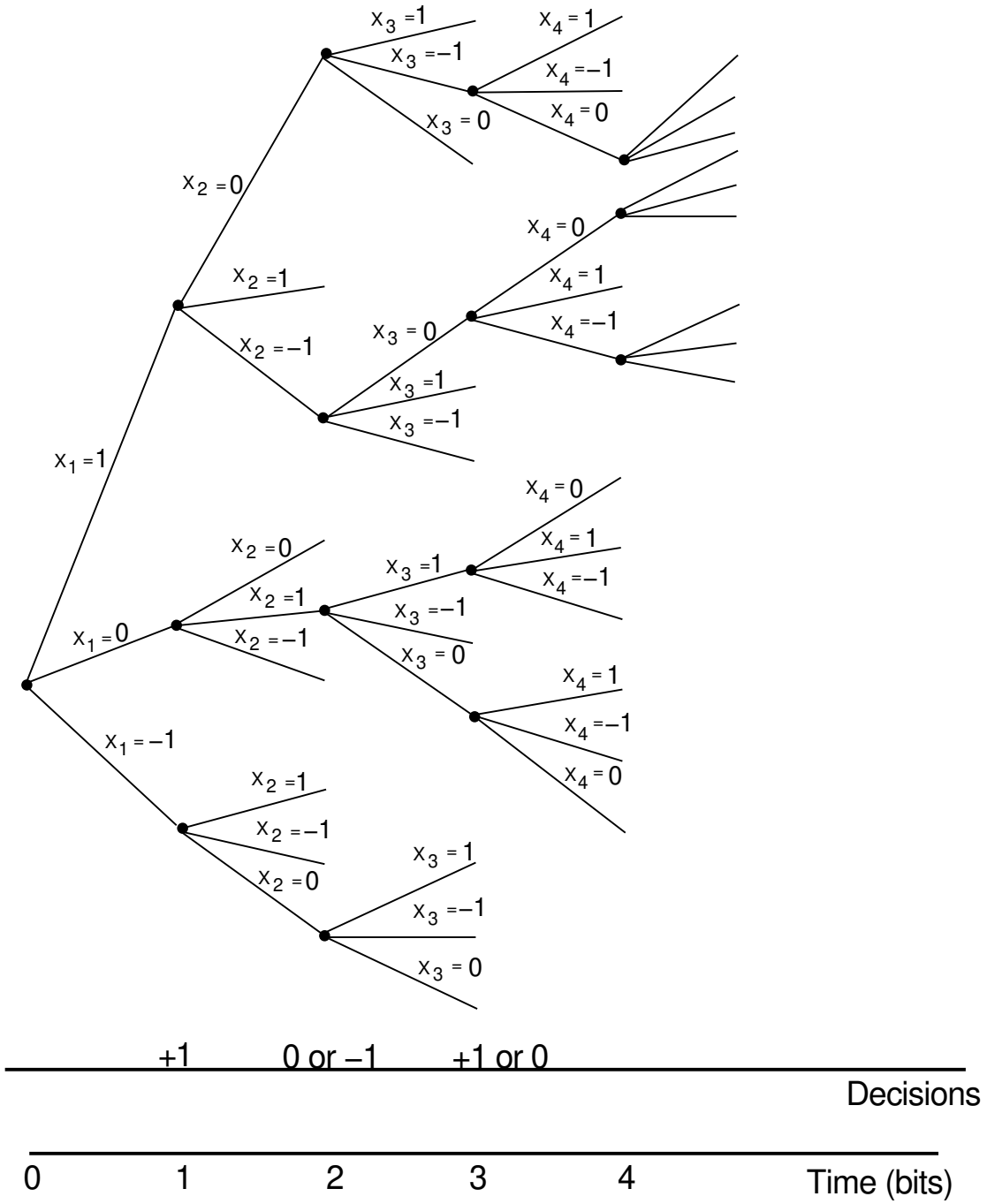
In the case of 3-level MDB, the receiver will select the least sum statistics for two timing points—i = 2 and i = 4.

Note that due to system limited memory (3 bits only), calculations in future will not use anything that happened before point 2, including point 2, as shown in Figure 4-26. But these have to be kept until only one sequence survives in the memory.

The tree diagram shown in Figure 4-27 illustrates the process. Between time moments 1 and 2, 9 statistics are kept in the memory. At point 2, 4 was selected, by 1 sequence at 2 lowest branches, and—unable to make up our minds for the highest branch—left two survivors. Therefore, 12 statistics were in memory between time moments 2 and 3,—at (after) moment 3, just 4 survived,—after moment 4, just 3 survived.

At some point only one survival point is left. That means a decision is made for that point. It is called survival path.

Figure 4–27: Tree Diagram



On the tree diagram at point i = 4:

1. Certain decisions about points i = 1 can be made. We claim that +1 was transmitted, and just its option has survived.

2. There is still some doubt in regards to time instants 2 and 3. At instant 2, it was or 0 or -1, at instant 3 it was either +1 or 0.

We will be able to make a decision about it a couple of steps later. Usually after 5 bits, all possible passes usually came into one survival and we can make decisions.

In general, to take advantage of using VA, the measurements of Yi must be pretty accurate. For instance, at 8-level quantization, 0.2 dB in performance is lost. With 2-level (binary), 2 dB is lost.

If we are going to use AV to fight the ISI too, the number of points in a branch becomes greater than 2; the number of calculations grows exponentially with the value of L.

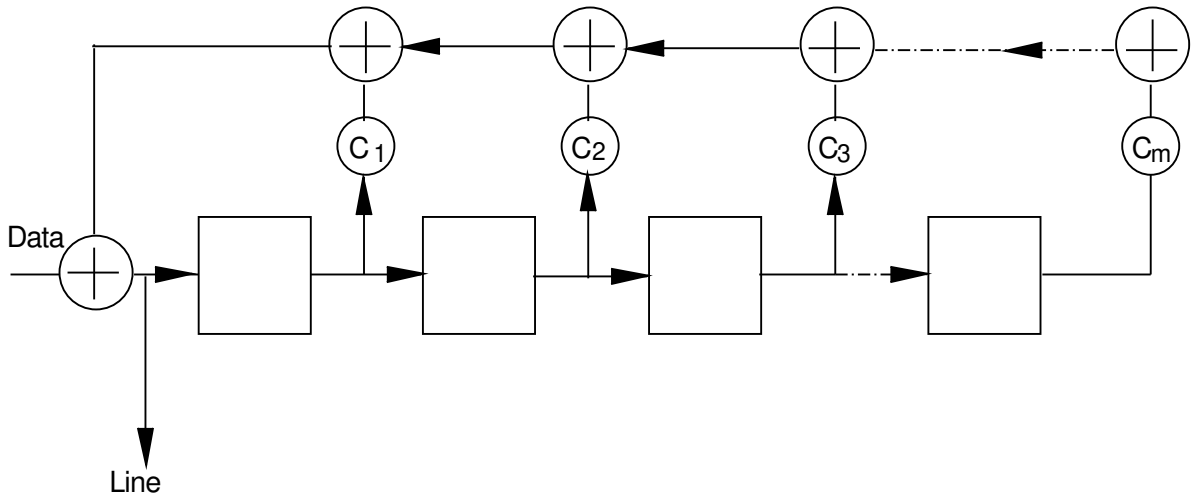
VA is successfully used in kbit frequency range. But for FDDI-related speeds, this algorithm represents quite a challenge.

4.3.4 SCRAMBLING

The purpose of scrambling is to eliminate periodic data patterns with short periods that contribute to emission in the frequency range beyond 30 MHz. The shortest in FDDI is 101010... with a 2-bit period. Actually, the scrambling increases the period of the input data.

The binary input data stream enters the scrambler feedback shift register where it is modulo-2 added to the pseudorandom patterns (feedback path), — self-synchronizing scrambler—producing the scrambler output shown in Figure 4-28.

Figure 4–28: General Basic Scrambler



The smallest period of scrambler output for the input sequence period M is, in general:

$T = M(2^N - 1)$

where N is the number of stages in the shift register.

For example, a 2-bit period of IDLEs (M = 2) with 7-bit shift (N = 7) register becomes 2 * 128 = 256 bit period, which corresponds to decreasing the frequency of periodic signal by 128 times.

In general, all scramblers can lock up if the long sequence of 0 bits quiets and all the elements of the scrambler shift register are 0 bits at the start of the long sequence of 0 bits. Actually, in this case, for the duration of 0 bits at the scrambler input there will be 0 bits generated at the output.

The consequences of the lock up on 0s is that the receiver PLL will eventually lose lock.

All scramblers with an even number of taps can lock up if a long sequence of 1s is at the the shift register input and all the elements of the scrambler shift register are 1 bits at the start of the long sequence of 1 bits.

These kinds of repeating patterns result in constant signal level, causing the clock extractor to lose its clock, or in a burst of high level electromagnetic emission.

To eliminate the periodic data at the scrambler output—scrambler lock up — a single-counter version of the scramblers, shown in Figures 4-29 and 4-30, is employed [22].

Figure 4–29: Single-Counter Descrambler

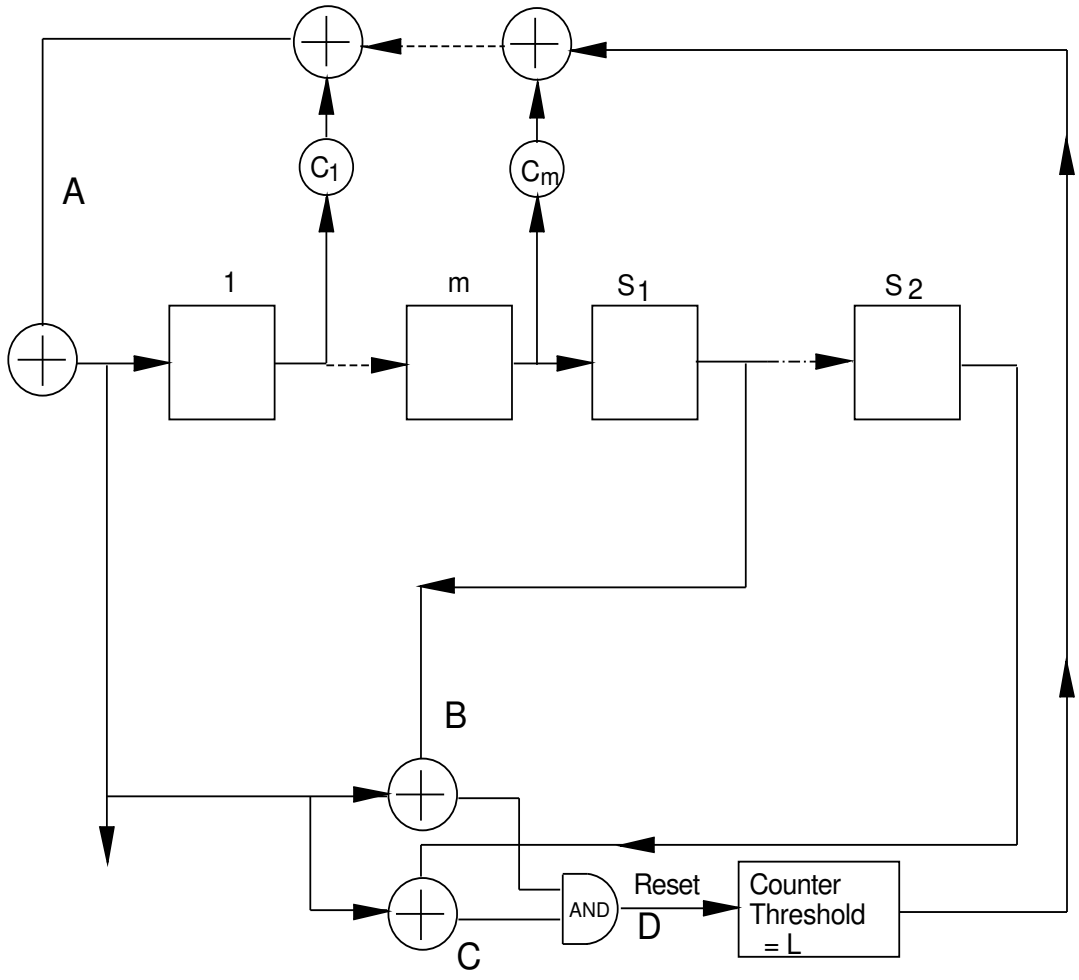
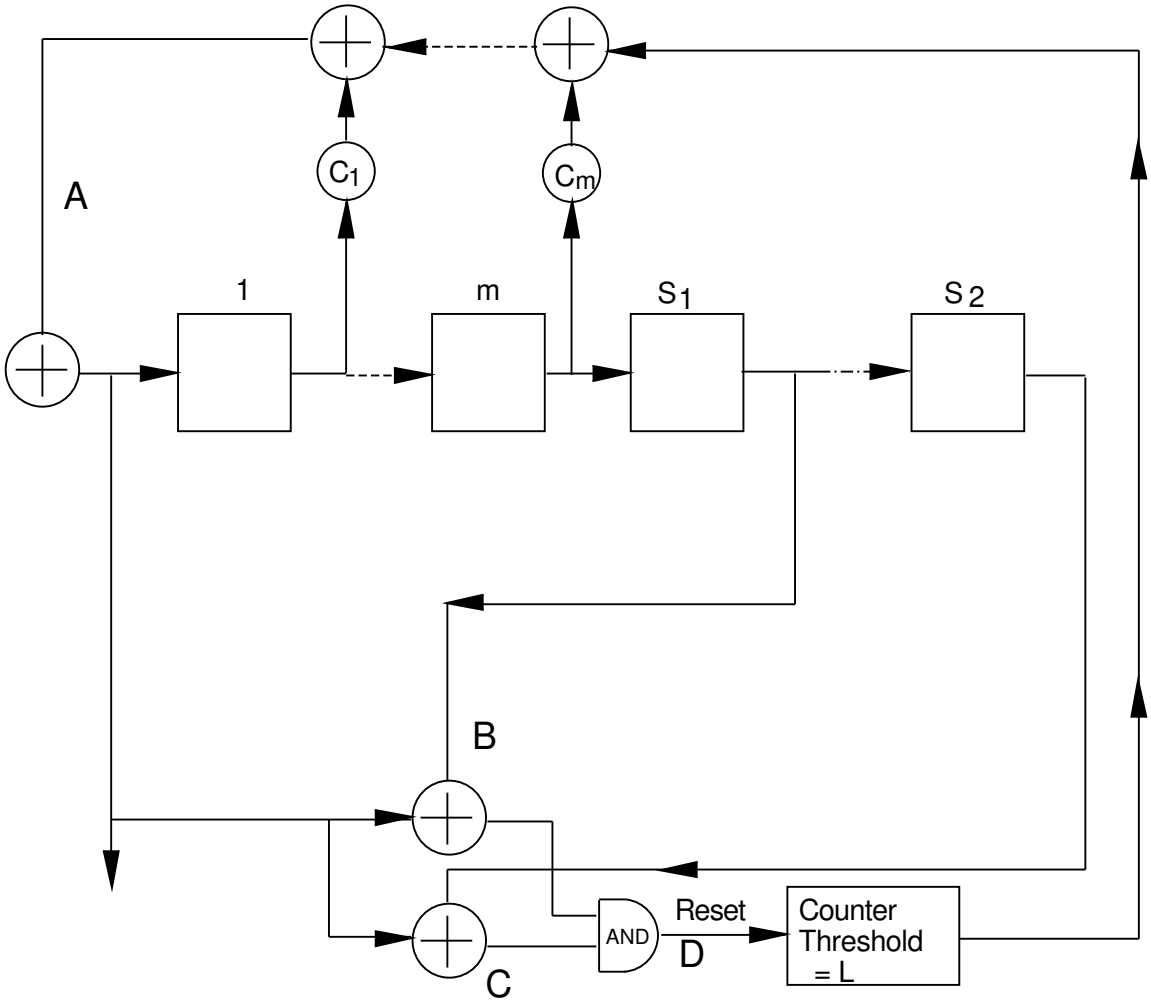


Figure 4–30: Single-Counter Scrambler



If incoming data contains periodic data such as one or both of the AND gate inputs are 0, the counter continues to count without being reset, eventually triggering the threshold. When the threshold is triggered, the state of the proper stage of the shift register is changed and periodicity is removed.

For example, if the scrambler has locked on 1s, then a single 0 bit is forced into the data stream entering the scrambler. This unlocks the scrambler and forces the corresponding bit into the receiver to be a 0, resulting in an error if that bit was originally a 1.

A single bit error will be detected and a frame will be re-transmitted. In general, locking up is an infrequent event and frame re-transmission can be tolerated [23].

A drawback of all the self-synchronizing scramblers is error propagation. When an erroneous bit enters the descrambler circuit, it causes error multiplication. If there are N taps in the descrambler, each isolated error in the received sequence will introduce N error counts.

Impulse noise environment can cause bursts of errors, which can result in two undesirable effects:

1. Long error bursts exceeding FCS ability

2. Frames created in correct format, but actually false

So, it seems important to select register length and the taps distribution which minimize these two factors.

An alternative is to use stream cipher, which spreads the signal spectrum but does not produce error multiplication.

Stream cipher is a modulo-2 addition of a pseudorandom sequence to data. The resultant data stream has the appearance and spectrum of essentially random data. This random data can be sent to a receiver, where the same pseudorandom sequence is again modulo-2 added. Because the modulo-2 addition is equivalent to subtraction, the original data stream is recovered.

The advantage of the stream cipher system is that a single error in the channel causes just one error at the receiver. The disadvantage is that the pseudorandom sequence at the receiver must be synchronized to the one at the transmitter. Some synchronization systems are proposed in [24].

4.3.5 SUMMARY

To provide sufficient margin against external noise, which is basicly unknown, it is worthwhile to have reasonably high transmit levels. The low bandwidth transmission systems enable a high transmission level without encountering FCC emission problems and, at the same time, result in the lowest possible signal attenuation on the line. Emission and susceptibility issues indicate desirability of moving the signal spectrum below 30 MHz.

Three approaches: 3B10, 32-CAP, and 4-pair MDB have the lowest spectral density above 30 MHz and consequently would allow much larger driving voltage.

Eight-level 3B10, 5- or 6-level 32-CAP, and 4-level MLT-4 most likely will require cross talk suppressors. In addition, MLT-4 requires dc coupling.

2B1Q, MLT-3 and 2-pair MDB with bandwidths of 62 MHz will require adaptive equalization and, perhaps, VA to meet FCC requirements at commercial distance. In addition, MLT-3 requires dc coupling.

Duobinary format (1 + D) provides close to MLT-3 spectrum, but does not have error propagation and provides a shorter run length.

In QAM-32, the frequency range of 50 MHz (25 MHz for each side band) and 5- or 6-levels would necessitate using a cross talk suppressor and, perhaps, adaptive equalization and scrambling.

Because of the lower modulation rate, spectral null at dc and near null at band edge frequencies, equalization for MDBs and 2B10 is considered less difficult than for MLT-3 [25].

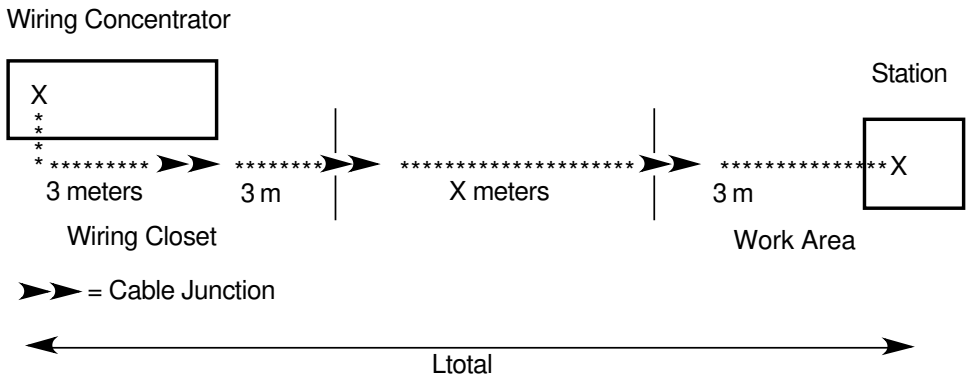
The cross talk supressor appears quite challenging because it includes multibit D/A, A/D and a significant amount of digital signal processing. It seems easier to implement it for MDB schemes because of the lower modulation rate.

Potentially, the best proposed modulation technique is CAP-32. But that requires a scrambler, in phase and quadrature filters, 8-bit D/A and A/D converters, adaptive filters for the receiver, and a cross talk suppressor.

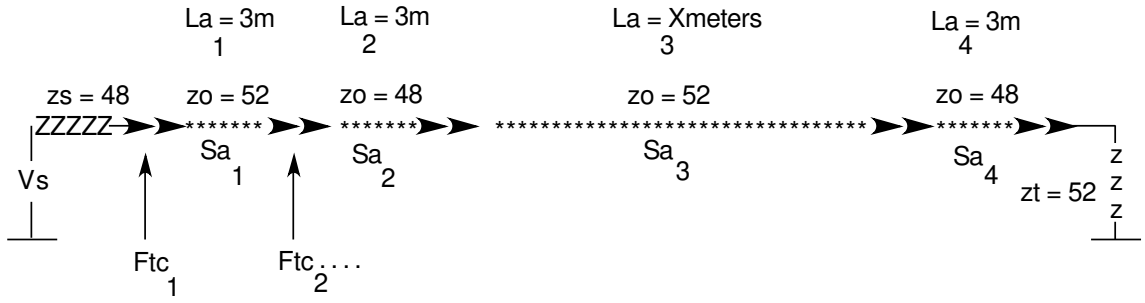
APPENDIX A REFLECTION LIMITED COAXIAL LINK: ACHIEVABLE DISTANCE

CROSS-CONNECT CONFIGURATION

Consider the following setup:



A worst-case electrical model for this setup is as follows:



Let the forward and reverse transmission coefficients of a segment n be given by [10]:

$$Ftc_n^{**} = \frac{2 * Z_l}{(Z_l + Z_o)}$$

$$Rtc_n = \frac{2 * Z_0}{(Z_l + Z_o)}$$

where:

Z_l = the effective load impedance
Z_o = characteristic impedance

Next let

$$Rc_n = \frac{Z_l - Z_o}{(Z_l + Z_o)}$$

be the reflection coefficient of segment n

Finally, let the attenuation of segment n, of length La, (measured at 62.5 MHz be defined as Sa. n

Therefore, $Sa_1 \ for \ La_1 = 3 \ meters^n$, is determined from:

$$Sa_1 = 10^{-(12/20*3/100)} = 0.959$$

where a loss of $-12dB/100$ meters @62.5 MHz is assumed.

Because the length of segment La_3 is related to the calculated value of Sa_3, one only needs to determine Sa to obtain this unknown segment. Sa_3 can be obtained by first determining the signal reflected back to the source from the local transmitter Vs defined as:

$$Vsr = Vs * Rc_1 + Vs * Ftc_1 * Sa_1^2 * Rc_2 * Rtc_1 + Vs * Ftc_1 * Sa_1^2 * Ftc_2 * Sa_2^2 * Rc_3 * Rtc_2 * Rtc_1$$

Note that the signal reflected back by segment 4 is small enough to be ignored.

and, the signal received from the far end transmitter is:

$$Vr = Vs_f * Rtc_5 * Sa_4 * Rtc_4 * Sa_3 * Rtc_3 * Sa_2 * Rtc_2 * Sa_1 * Rtc_1$$

where: Vs_f = the far end transmitter signal.

Allowing for a 6 dB signal/reflected noise ratio and -35 dB local transmitter rejection then:

$$Vr = 2 * (Vsr + 0.018Vs)$$

Solving for Sa_3 one obtains:

$$Sa_3 = 2 * \frac{\left\{Rc_1 + Ftc_1 * Sa_1^2 * Rc_2 * Rtc_1 + Ftc_1 * Sa_1^2 * Ftc_2 * Sa_2^2 * Rc_3 * Rtc_2 * Rtc_1 + 0.018\right\}}{\{Rtc_5 * Sa_4 * Rtc_4 * Rtc_3 * Sa_2 * Rtc_2 * Sa_1 * Rtc_1\}} * \frac{Vs}{Vsf}$$

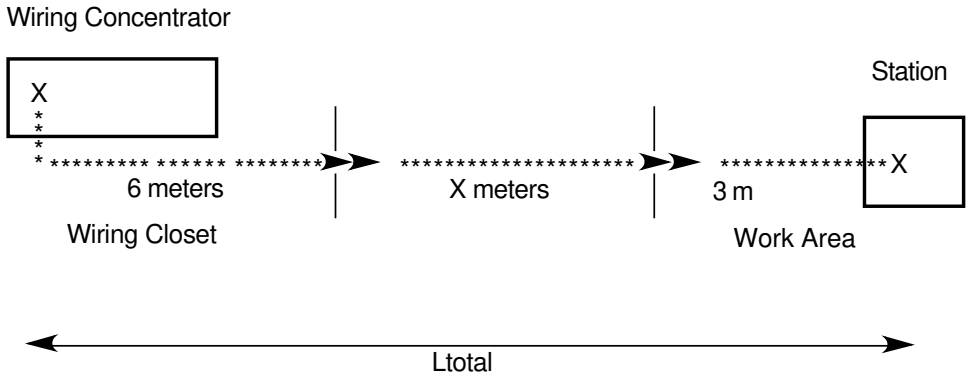
Upon substitution of values* for the above electrical model one obtains:

$Sa_3 = 0.430$ or an equivalent of 61 meters

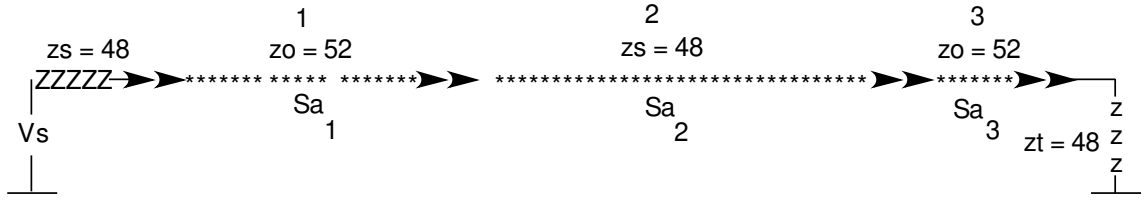
Resulting in a total length for this particular case of:

$$Ltotal = 61 + 9 = \underline{70 \ meters}$$

Standard Configuration



A worst-case electrical model for the above is:



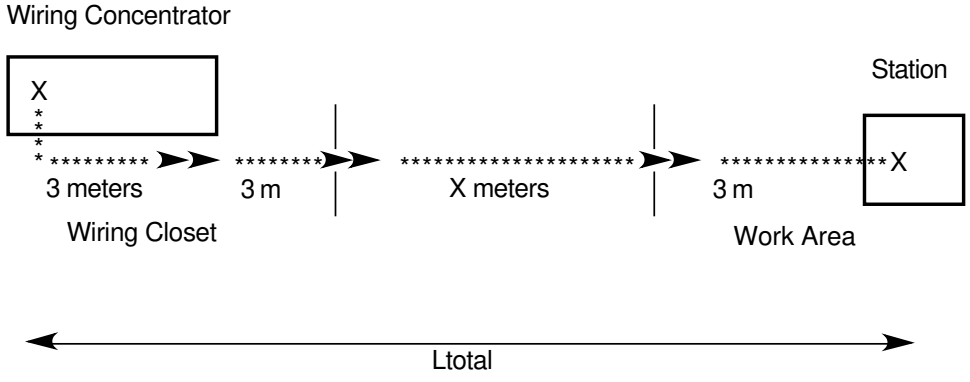
Using an analysis similar to the cross-connect configuration, one obtains:

$$Sa_2 = 2 * \frac{\left\{Rc_1 + Ftc_1 * Sa_1^2 * Rc_2 * Rtc_1 + 0.018\right\}}{\{Rtc * Sa * Rtc * Rtc * Sa * Rtc*\}} * \frac{Vs}{Vs_f}$$

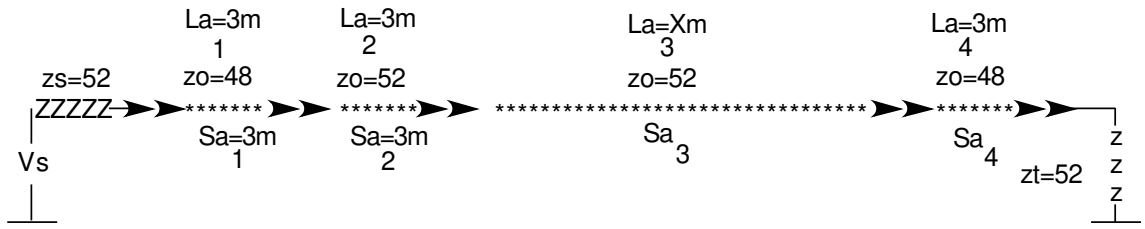
and $\underline{Ltotal = 97.3\ meters}$

Typical cross-connect configuration

Finally, consider the following cross-connect configuration where the cable impedance variations are not worst-case.



The electrical model for the previous figure is shown here.



Similar to the first analysis, the results are:

$$Sa_3 = 2 * \frac{\left\{ Rc_1 + Ftc_1 * Sa_1^2 * Rc_2 * Rtc_1 + Ftc_1 * Sa_1^2 * Ftc_2 * Sa_2^2 * Rc_3 * Rtc_2 * Rtc_1 + 0.018 \right\}}{\{Rtc_5 * Sa_4 * Rtc_4 * Rtc_3 * Sa_2 * Rtc_2 * Sa_1 * Rtc_1\}} * \frac{Vs}{Vs_f}$$

Upon substitution, $Sa_3 = 0.305$

and, $\underline{Ltotal = 95\ meters}$

APPENDIX B UTP: ACHIEVABLE DISTANCE

Because UTP with MDB does not have BLW and reflections are not essential, the expression (3) of subhead 1.13 can be rewritten in the form:

$$\frac{SK}{n - 1} \geq SX + D$$

where
n = Number of transmitted levels
K = Transmission coefficient for signal
X = Transmission coefficient for the NEXT
D = Induced noise (rather than cross talk) p-p (V)
S = Transmitted signal level (V)
We can find

$$K \geq (n - 1)\left[X + \frac{(n - 1)D}{S}\right]$$

All values are peak-to-peak.

Using the last expression, we can calculate the achievable distance.

The number of ports is 12.

Allowable (by FCC regulations) driving voltage:

$$S_2 = 0.35/(\sqrt{12}) = 0.1 \ V = 100 \ mV \quad for \ \ 125MHz$$

$$S_3 = 0.41/(\sqrt{12}) = 0.12 \ V = 120 \ mV \quad for \ 62.5MHz$$

$$S_7 = 0.61/(\sqrt{12}) = 0.18 \ V = 180 \ mV \quad for \ \ 42MHz$$

$$S_{15} = 0.87/(\sqrt{12}) = 0.25 \ V = 250 \ mV \quad for \ \ 31MHz$$

Externally induced noise is assumed $D = 10mV$.

Cross talk obtained from NT data:

$$X_2 = (-30dB) = 0.03 \qquad at \ 62MHz$$

$$X_3 = (-35dB) = 0.018 \qquad at \ 31MHz$$

$$X_7 = (-39dB) = 0.011 \qquad at \ 21MHz$$

$$X_{15} = (-41dB) = 0.009 \qquad at \ 16MHz$$

Northern Telecom data (average)

Now we obtain the signal transmission coefficient:

$K_2 = 2(0.03 + \frac{20}{100}) = 0.46$ allowable losses $1/K_2$ — $6.74dB$

$K_3 = 2(0.018 + \frac{20}{120}) = 0.37$ allowable losses $1/K_3$ — $8.64dB$

$K_7 = 6(0.011 + \frac{60}{180}) = 2.1$ allowable losses $1/K_7$ — negative db (no distance available)

$K_{15} = 14(0.009 + \frac{140}{250}) = 8$ allowable losses $1/K_{15}$ — negative db (no distance available)

Now we find signal attenuation per meter from NT data:

$K_2 - - - 0.31dB/m\ at\ 62MHz\ (31dB\ per\ 300ft)$
$K_3 - - - 0.18dB/m\ at\ 31MHz\ (18dB\ per\ 300ft)$
$K_7 - - - 0.14dB/m\ at\ 21MHz\ (14dB\ per\ 300ft)$
$K_{15} - - - 0.12dB/m\ at\ 16MHz\ (12dB\ per\ 300ft)$

Northern Telecom data (average)

Thus the achievable distance is

2 level system $6.74/0.31 = 21.7_m$
3 level system $8.64/0.18 = 48_m$
7 level system no distance available
15 level system no distance available (distance between levels less than noise)

Effect of Induced Noise and Cross Talk

The distance is a strong function of induced noise. For instance, if we decrease $10mV$ to $5mV$ we'll have

$K_{3'} = 2(0.018 + \frac{10}{120}) = 0.2$

allowable losses $1/K_3$ — $14dB$

and achievable distance for a 3-level system is $14/0.18 = 77_m$ which corresponds to increase by 60%.

In this model, based on measured data the external noise ($3V/m$) dominates. Dealing with an environment with low external noise, the distance will be in direct proportion to cross talk attenuation.

If cross talk dominates the achievable distance is in direct proportion to the cross talk attenuation: change in the cross talk by 6 dB will result in changing in distance by $6/0.18 = 33_m$ (3-level, 2-pair of NT cable).

Crescendo's 8 ports provide additional distance compared with 12 ports of

$$\sqrt{(12/8)} = 1.22$$

this 1.22 results in allowable voltage of $120 * 1.22 = 146.4$ and

$K_{3in.} = 2(0.018 + \frac{20}{146.4}) = 0.31$ allowable losses $1/K_{3in.}$ — $10.2dB$

A new distance is $10.2/0.18 = 57_m$

For example, additional distance of $57 - 48 = 9_m$ or 19%.

APPENDIX C BIBLIOGRAPHY

1. Tsao, David. *FDDI: Chapter 2*. Data Communications, September 21, 1991, p 59.
2. Fontolliet, P.G. *Telecommunications Systems*. Artech House, 1986.
3. Freeman, Roger L.*Telecommunication System Engineering*.John Wiley and Sons, 1989, p 267.
4. Shamai, Shlomo. *On the Capacity of Twisted Pair*. IEEE Transactions on Communications, Vol. 38, No. 3, March, 1990, pp 368 - 383.
5. Johnson, Don C. *Impulse Noise*. June, 1988. Contribution to 10 base-T, p.25.
6. Jain, R. *Error Characteristics of Fiber Distributed Data Interface (FDDI)*. IEEE Transactions on Communications, Vol 38, No. 8, August, 1990.
7. Yang, Henry and Jerry Hutchinson and Paul Koning. *FDDI Data Link Functional Specification*. Digital specification. 18 July 1990.
8. *Data Communications*. 21 September 1990, pp 41 - 50.
9. *EIA/TIA Cable Specifications for UTP* 9 September 1991.
10. Fitzgerald, John. *125-Mbit STP Link Transceiver Specification*. Digital specification. September 14, 1990.
11. McDonald, Elaine M., and Jonathan Hill and Len Irwin. *FDDI Copper DVT Report*. August 1, 1991.
12. Hill, Jonathan. *Transmitter DVT Report for DEFCN-TA and DEFZA-CA in Coaxial Mode*. NSE Report, January 15, 1992.
13. Fitzgerald, John. *125-Mbit STP Link Transceiver Specification*. Digital specification. March 1, 1991.
14. Hill, Jonathan. *Transmitter DVT Report for DEFCN-SA and DEFZA-CA in STP Mode*. NSE Report, January 13, 1992.
15. Proakis, John G. *Digital Communications*. McGraw-Hill, 1989 p. 276.
16. Cook, John. Contribution to ANSI X3T9.5 October, 1991.
17. Boakue, K.A., et al. IEEE 802.9. Contribution 802.9 - 90/9. January 15, 1990.
18. Ginzburg, Simon, and William Mallard and David Newman. *FDDI Over Unshielded Twisted Pairs*. Procedures of the 15th Conference on Local Computer Networks. 1990, pp 395 - 398.
19. Kim, D.Y., et al. *Run Length-Limited Variants of Duobinary and Modified Duobinary*. IEEE Transactions on Communications, Vol. com-35 N2 February, 1987.
20. Mazzola, Mario. Presentation to ANSI X3T9.5, TP/PMD Working Group. 1990
21. Silicon Systems. Presentation to ANSI X3T9.5 TP/PMD Working Group, San Diego, CA. December 9, 1991.

22. Savage, J.E. *Some Simple Self-Synchronizing Digital Data Scramblers*. The Bell System Technical Journal, February 1967, pp. 449 - 457.

23. Spratt, Michael. *The Use of Scramblers with an Antilocking Circuit*. Contribution to ANSI X3T9.5, November, 1991.

24. Proudler, G. *Stream Cypher for CDDI*. Contribution to ANSI X3T9.5, November, 1991.

25. Ungerboek, G. *Comparison of MLT-3, QPR-IV, and Schemes Beyond QPR-IV*. Presentation to ANSI X3T9.5 TP/PMD Working Group. San Diego, CA. December 9, 1991.

APPENDIX D RELATED DOCUMENTS

D.1 Other Digital Documents

Document Number	Document Title
EL-EN487-00	*STP Rules, Data Requirements, Algorithm, and Strategy*
EL-EN529-00	*FDDI (Fiber Dtstributed Data Interface) Physical Channel Components Test Procedures*
EL-EN529-01	*DEC FDDI Physical Channel: A Technical Report*

```
+---------------------------------------------------------------------+
|                          READER COMMENTS                            |
|                                                                     |
| Your  comments  and suggestions  will help  Standards and Methods   |
| Control improve their services and documents.                       |
+---------------------------------------------------------------------+
```

Did you request this document? _____ If so, did it arrive within a satisfactory period of time? _____ Please comment.

What are your impressions of this document? Consider format, organization, completeness, readability, and illustrations.

————————————**FOLD ON THIS LINE**———————————————————————— -

Did you find technical or clerical errors in this document? If so, please specify the page number(s) and the error(s).

Are the instructions for the update package clear? _____
Was an index available? _____ If not, is one needed? _____
Do you have other suggestions for improving this document?

The following information is optional:

Name _____ Mailstop _____

Department _____ Node _____

Send your comments to JOKUR::PROJECTS, or fold, staple, and send this page through interoffice mail to:

```
+------------------------------+
|       READERS' COMMENTS      |
| STANDARDS AND METHODS CONTROL |
|           NRO4/D4            |
+------------------------------+
```